Amendments to the Specification:

Page 1, line 10, please amend the following paragraphs:

~~COMPUTER PROGRAM LISTING APPENDIX~~

~~The computer program listing appendix attached hereto consists of two (2) identical compact disks, copy 1 and copy 2, each containing a listing of the software code for embodiments of components of this invention. Each compact disk contains the following files (date and time of creation, size in bytes, filename):~~

~~Directory of D:\~~

~~05/30/2003~~	~~09:09 AM~~	~~1,188~~	~~0711115B.txt~~
~~05/30/2003~~	~~09:11 AM~~	~~7,671~~	~~0711115C.txt~~
~~05/30/2003~~	~~09:12 AM~~	~~1,021~~	~~0711115D.txt~~
~~05/30/2003~~	~~09:18 AM~~	~~1,361~~	~~0711115E.txt~~
~~05/30/2003~~	~~09:19 AM~~	~~335~~	~~0711115F.txt~~
~~05/30/2003~~	~~09:20 AM~~	~~649~~	~~0711115G.txt~~
~~05/30/2003~~	~~09:08 AM~~	~~3,989~~	~~071115A.txt~~
~~05/30/2003~~	~~09:05 AM~~	~~38,253 071119.txt~~	
	~~8 File(s)~~	~~54,467 bytes~~	
	~~0 Dir(s)~~	~~0 bytes free~~	

~~The contents of the compact disk are a part of the present disclosure, and are incorporated by reference herein in their entireties.~~

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Page 4, line 10, please amend the following paragraph:

There are a number of causes of audio distortion ~~which~~ that involve the structure and operation of the voice coil transducer 300. At high signal levels, voice coil transducers become very distorting. This distortion is largely caused by the nonlinearities in the coil motor factor, in the restoring force of the coil/diaphragm assembly suspension, and the impedance of the coil. Other nonlinear effects also contribute to the distortion. Nonlinear effects are an intrinsic part of the design of voice coil transducers.

Page 5, line 7, please amend the following paragraph:

FIG. 4 represents data for actual large signal (LS) parameters of a transducer from a small desktop stereo system, model name: Spin70, manufactured by Labtec. The large signal parameters shown in FIG. 4 were obtained using a commercially available laser metrology system (Klippel GMBH). The magnitude of Bl is shown by curve 401 as a function of the displacement x of the coil/diaphragm assembly from the no-signal equilibrium position, which is indicated in FIG. 4 by a zero on the horizontal axis; at that position, no elastic restoring force is applied to the -coil/diaphragm assembly. The unit for Bl is $Newton\,/\,Ampere$ (or $N\,/\,A$). The highly non-constant nature of the Bl factors of commercial voice coil transducers is recognized in the current art. As the audio signal increases in magnitude, the coil tends to move away from the region of maximal static magnetic field, and the motor factor decreases, thus effecting a less uniform coil movement and distorting the sound wave.

Page 5, line 23, please amend the following paragraph:

Referring to FIG. 3, as pointed out above, the cone suspension is axially symmetric and typically includes two parts: a corrugated suspension near the coil, typically referred to as the spider 307, and the surround 308 connecting the large end of cone 302 to the frame 301 of the speaker. These two suspensions together act as an effective spring, which provides a restoring force to the coil/diaphragm assembly and determines the equilibrium position of the assembly to which it relaxes when not being driven. This effective spring restoring force is again a highly non-constant function of coil/cone axial position x; that is to say, the effective spring stiffness varies significantly as a function of x. In FIG. 4 curve 402 shows a plot of K, the spring stiffness, as a function of x for the speaker transducer mentioned above. Spring stiffness K is expressed in units of $N\,/\,mm$ (-i.e. Newtons per millimeter).

Page 6, line 13, please amend the following paragraph:

———where m is the mass of the assembly plus a correction for the mass of air being moved; R_{ms} represents the effective drag coefficient experienced by the -assembly, mainly due to air back pressure and suspension friction; $K(x)$ is the position dependent effective spring stiffness due to the elastic suspension; $Bl(x)$ is the position dependent motor factor; and $i(t)$ is the time dependent voice-coil current, which responds to the input audio signal and constitutes the control variable. These terms are related to the industry standard linear model (small signal) parameters – namely, the Thiele-Small parameters, which are as follows:

Page 7, line 4, please amend the following paragraphs:

The second order differential equation (1) would be straightforward to solve, but for the nonlinearities in the elastic restoring force and in the motor force terms; these nonlinearities stem from the x dependence of $K(x)$ and $Bl(x)$, and- they preclude a closed-form analytical solution in the general case. Although approximations can be made, it is difficult to predict the response of a system under all conditions, and thus to create a robust control system.

Further nonlinearities arise due to other electrodynamical effects caused by the application of the audio signal to the transducer voice-coil. Typically, current is supplied to the coil by converting the audio information into a voltage, *V(t)*, which is imposed across the terminals of the voice coil. However, the resulting coil current varies both out of phase and nonlinearly with this voltage. The phase lag arises both because the voice coil's effective impedance has a reactive component, and because the electromechanical transduction of the coil current into coil motion through the -static magnetic field induces a Back–ElectroMotive Force *(BEMF)* voltage term in the coil circuit.

Page 7, line 30, please amend the following paragraph:

—~~Where~~ where the BEMF is represented by the second term on the left hand side (a product of $Bl(x)$ and coil velocity). The Ohmic resistance of the coil is R_e. The coil's effective inductance, $L_e(x)$, is a function of x because it depends upon the instantaneous position of the coil relative to the magnetic pole structure and its airgap. In FIG. 4 curve 403 shows a typical plot of the position dependence of coil inductance $L_e(x)$ at low audio frequencies. The units of L_e are mH (milli-Henries), and the values of L_e shown in curve 403 have been multiplied by a factor 10 to render the graph more readable.

Page 9, line 7: equation (4)

$$\cancel{\phi(x) = -\frac{K(x)}{m}x - \frac{R_m}{m}\dot{x}}\ \phi(x) = -\frac{K(x)}{m}x - \frac{R_{ms}}{m}\dot{x} \tag{4}$$

Page 10, line 10, please amend the following paragraph:

Accordingly, there are several factors described above that significantly affect the ability to provide -accurate sound from a conventional audio reproduction system. Some of the issues can be addressed by improving the circuitry through digital means; but even with the digital circuitry to handle the signal shaping, the transducer itself has significant nonlinearities that can never be addressed adequately by shaping the input signal to the transducer. Therefore, what is needed is a system that controls the transducer in such a manner that optimum linear sound is provided. Such a system should also be easy to implement, cost effective, and easily adaptable to existing systems. The present invention provides a control system for a transducer to provide linear sound, and the present invention also provides an integrated audio reproduction system.

Page 13, line 18, please amend the following paragraph:

FIG. 10 is a block diagram of the feedback linearization process using the control law given by equation ~~(38)~~(40), which provides transduction corrections along with a linear spring constant (suspension stiffness) ~~which~~ that is electronically added;

Page 14, line 2, please amend the following paragraph:

FIG. 14 illustrates Power Spectrum Distribution simulation curves showing the effect of the transduction corrections (spring stiffness and motor factor correction) upon harmonic distortion for a single- $100\,Hz$ tone input, both with and without BEMF and nonlinear inductance in the physical model of the Labtec Spin 70 transducer;

Page 14, line 11, please amend the following paragraph:

FIG. 16 illustrates simulated waveforms of the coil/diaphragm axial position versus time in the presence of a single-tone excitation, both with and without electronically restored effective spring stiffness, showing that without such restoration the cone may drift from its equilibrium position and reach its limit of excursion;

Page 14, line 20, please amend the following paragraph:

FIG. 18 is a graph of the simulated phase lag between coil voltage and coil current as a function of audio frequency at low frequencies, which is almost entirely due to BEMF;

Page 15, line 10, please amend the following paragraph:

FIG. 24 illustrates a plot of suspension stiffness K in Newtons/mm together with a plot of Bl in Newtons/amp, both of which are plotted against L_e for the same Labtec Spin 70 transducer data;

Page 15, line 16, please amend the following paragraph:

FIG. 26 shows a curve ~~which~~ that illustrates the variation of L_e with position at 43kHz for a Labtec Spin70 transducer;

Page 15, line 24, please amend the following paragraph:

FIG. 31 is a block diagram for a circuit ~~which~~that, together with parameter estimation, measures transducer coil inductance via a supersonic probe tone and reference RL circuit;

Page 15, line 30, please amend the following paragraph:

FIG. 33 shows a curve ~~which~~ that illustrates the variation of $C_{parasitic}$ with V_{coil} for driven measurement; $C_{parasitic}$ is measured in arbitrary units obtained using the method described in Detail 12;

Page 16, line 9, please amend the following paragraph:

FIG. 37 is a cross-sectional view of a speaker transducer ~~which~~ that includes an IR-LED diode and an associated PIN diode, mounted on the back side of an audio transducer of the type shown in FIG. 3, as part of an optical position detection system;

Page 16, line 27, please amend the following paragraph:

FIG. 43 shows the details of HW and ISR operations for the S calibration in step 11504 of FIG. ~~115~~42;

Page 17, line 1, please amend the following paragraphs:

FIG. 46 illustrates the operations of process 11203 of FIG. 45 that are spawned as a result of enabling sampling clock and ISR in step 11202 of FIG. ~~112~~45;

FIG. 47 shows a flow diagram of the ISR 11303 of FIG. ~~113~~46;

Page 18, line 16, please amend the following paragraph:

FIG. 66 shows a low-frequency portion of the same power spectrum distribution shown in FIG. 65, displaying multiple harmonics of the $60\,Hz$ tone, with spectra depicted both with and without correction.

DETAILED DESCRIPTION OF THE EMBODIMENT(S)

Page 19, line 1, please amend the following paragraph:

An enabling invention in the area of control engineering 501 was the linearization method for dynamical equations 504 used in modeling physical systems to be controlled, such as actuators and transducers. This method relies on finding the control equation for the non-linear part of the dynamical equation and substituting this into the full equation. The application of this method to a second order differential equation 505 shows that a non-linear second order ordinary differential equation can be linearized by solving the control equation for the non-linear first order differential equation, provided the second order and first order differential terms are linear. This is a general method for linearizing such differential equations, and covers the application to the control of all actuators and transducer systems that can be modeled in full, or in part, by such an equation. The application of the linearizing method 505 to an equation with nonlinearities dependent on one state variable 506 shows that only one state variable is required for linearization. The application of 506 relies on positional sensing. That is to say, neither the velocity, nor the acceleration, nor the instantaneous driving force state variables are required in order to linearize the process. Position dependent sensing and feedback linearization can be used with many classes of ~~a~~non-linear motors and actuators.

Page 19, line 25, please amend the following paragraphs:

In the present work it was discovered that there are multiple processes in a sound reproduction system, that each process can influence the performance of other processes, that each process has non-linearities that must be considered in the design of a control ~~paradigm~~loop, and that each control ~~paradigm~~ loop must have a sufficient number of state measurements which must be measured with sufficient discrimination against noise and with sufficient speed to control the process. ~~It was further discovered that control of one process must be stable in the presence of other processes.~~

Control of multiple processes with multiple control ~~paradigms~~ loops can be ~~affected~~ effected if the criteria for sufficiency is met for each control ~~paradigm~~loop. It has been

discovered that for the correction of non-linear transduction a necessary condition for control is a positional state measurement, in distinction to the motional measurements of prior art. The positional state measurement must be of sufficiently low-_noise and latency and of sufficiently high ~~speed, or~~ bandwidth~~,~~ to effect the control while not adding unacceptable noise to, nor engendering instability in, the sound output. Multiple positional measurements can be used to estimate the positional state for the purpose of transducer linearization.

Page 20, line 30, please amend the following paragraph:

The method and system comprise providing a model of at least a portion of the audio transducer system and utilizing a control engineering technique in the time domain to control an output of the audio transducer system based upon the model. In the present invention a method to determine, in real-_time, the nonlinear parameters of the transducer from measurement of internal state parameters of the transducer is provided. In particular the electrical properties of the voice coil can be used as a measure of positional state and a predictor of the major non-linearities of the transducer. "Real-_time" in this context means with sufficiently ~~bandwidth~~ low latency to effect control.

Page 21, line 12, please amend the following paragraphs:

It has been discovered that in an audio reproduction system, the overall process of converting audio information into sound can be considered as consisting of three processes. First, conditioning of the audio signal to produce the transducer drive signal; second, the transduction of the drive signal into a diaphragm motion moving an air mass; and third, the conditioning of the moving air mass to provide an output sound. Thus, an audio transducer can be defined as: signal conditioning/transduction/sound conditioning. FIG. 6 illustrates a block diagram of audio reproduction system 1100 in accordance with these processes. As is seen, a signal conditioning process 1102 takes an audio signal 1101 (digital or analog) and performs signal conversion, amplification, filtering and frequency partitioning to provide a drive signal 1103. The drive signal 1103 is provided to a transduction process 1104. The transduction process 1104 typically utilizes a plurality of transducers, and results in diaphragm motion 1105, which drives an air load. A sound conditioning process 1106, which may include effects from a speaker enclosure and an extended audio environment, acts on the air load driven by the diaphragm motion 1105 to provide the perceived sound 1107.

Distorting factors due to nonlinear effects influence all of these processes. These factors arise in the relationship between the audio signal as a voltage and the drive current in the coil

(transconductance), and in the electro-magneto-mechanical (henceforth abbreviated "electromechanical") effects involving the moving-coil motor. Nonlinear effects resulting from sound conditioning are much smaller in normal operating conditions, and are thus neglected in the physical model described in this section, and in the control model based upon it and described in Detail 2. But these nonlinear acoustical effects, along with other higher-order effects described and then neglected in this section, can in principle also be linearized, via separate control ~~laws~~ loops according to the 'modular' approach to linearization disclosed as part of this invention.

All of the effects mentioned above vary with time and circumstances. They are nonlinear and thus distort the sound wave shape, in both amplitude and phase, relative to the input audio information. Furthermore, due to the inherently bi-directional nature of the transconductance and the electromechanical transduction, and of the coupling between them, distortions in any one process ~~are mirrored in~~can affect any of the other processes. Most importantly, it is the nonlinearities inherent in the electromechanical transduction ~~which~~ that make the linearization and control of the overall process very difficult in prior art.

Page 23, line 8, please amend the following paragraph:

There are well-recognized nonlinearities in the drive current as a function of voltage, caused by the dependence of effective coil impedance and of the motor's BEMF upon coil position relative to the magnet assembly. The effective spring stiffness of the coil/diaphragm assembly, likewise dependent on coil position, as is the motor factor, result in well-recognized sources of nonlinearity. Additionally, more gradual changes of coil impedance due to Ohmic and environmental heating cause the drive-current response to vary over time. All these effects cause power and frequency dependent distortions of the audio signal.

Page 24, line 9, please amend the following paragraph:

The three processes can be described by a mathematical model, comprising a system of coupled equations specifying the rate of change (evolution) of each of a complete set of state variables, such as coil current and coil position, at any given time, in terms of the state vector at the same and all previous times. Such equations are termed "integro-differential equations", and are nonlinear in the case at hand. In the prior art, the model equations are usually approximated as having no "memory", in the sense that the rates of change of state variables are taken to be wholly determined by (generally nonlinear functions of) state variables at the same instant of time; such memory-less evolution equations are simply termed "differential equations". ~~The~~

~~mathematical model according to the present invention, however, includes memory effects, as it has been discovered that they cannot, in general, be entirely neglected in modeling the audio reproduction system.~~

Page 24, line 32, please amend the following paragraph:

A nonlinear process can be very complex, and the number of terms kept in the evolution equations, as well as the decision whether or not to include memory effects, and if so which ones, can vary depending on the degree of approximation required in the control methodology. In the explanation ~~which~~ that follows, it will be seen that simplifying the approximations to the most basic mechanisms of the three processes yields several coupled "ordinary" nonlinear differential equations. Anyone skilled in the art will appreciate that using approximations is a compromise, and that beyond a certain point, enlarging or truncating the list of modeled effects does not alter the fundamentals of the invention.

Page 27, line 21, please amend the following paragraphs:

Most of the parameters and parameterized functions appearing in equations (6) through (~~11~~10), specifically R_e, R_{ms}, $Bl(x)$, $K(x)$, $h(t)$ and the functions g_1 through g_4, depend on temperature, which is assumed to vary slowly as compared with timescales characterizing audio response. For the approximation to be fully self-consistent, the acoustic-load part of R_{ms} should actually be replaced with a memory term related to $h(t)$; the fact that a constant R_{ms} is instead used in equation (9) is a further, non-essential approximation.

The time integrals in equations (7)~~,~~ and (8) encode memory effects due to eddy currents, while the integral in the pressure equation (10) encodes memory effects due to acoustic reflections and dispersion. All of these integrals represent the dependence of the rate of change of state variables at any given time, upon the history (past values) of those same state variables. Although effects from the infinitely remote past are in principle included in these integrals, in practice the memory of past positions and currents fades eventually, because the audio signal is band limited.

It has been found that, while the memory effects encoded in equations (7), (8)~~,~~ and (10) are important for modeling the dynamics of an audio reproduction system, they are ~~second-order~~of secondary importance in the context of ~~adistortion-~~ distortion-correction controller.

The spectral contributions to the dynamic coil excursion $x(t)$ are dominated by low frequencies, a fact well recognized in prior art. In consequence, it is often a reasonable approximation to replace the delayed positions $x(\tau)$, $x(\tau_1)$ and $x(\tau_2)$ in the memory integrals

of equations (7)-(8) with low-order Taylor expansions about the present time (i.e. about $\tau = t$, $\tau_1 = t$ and, $\tau_2 = t$ respectively.). In this way, positional memory effects are neglected,; while the more important memory effects involving delayed response to current and velocity, are still included. If this further approximation is implemented, and terms quadratic and higher in coil velocity are neglected, the electromechanical and elastic parts of the above system of evolution equations, equations (6) through (9), simplify to the following form.

The coil-circuit equation (governing the transconductance component of the signal conditioning process) becomes:

$$V_{coil}(t) = R_e\, i(t) + \dot{x}(t)\,\Phi_{dynamic}(t) + V_{efield}(t) \tag{11}$$

~~Where~~ where now $\Phi_{dynamic}(t)$ and $V_{efield}(t)$ simplify to

$$\begin{aligned}\Phi_{dynamic}(t) &= Bl(x(t)) + \int_{-\infty}^{t} d\tau\; g_1(t-\tau, x(t))\, i(\tau) + \\ &+ \int_{-\infty}^{t} d\tau_1 \int_{-\infty}^{\tau_1} d\tau_2\; g_2^{(0)}(t-\tau_1, t-\tau_2, x(t))\; i(\tau_1) i(\tau_2)\end{aligned} \tag{12}$$

—and

$$\begin{aligned}V_{efield}(t) &= \int_{-\infty}^{t} d\tau\; g_3(t-\tau, x(t))\, i(\tau) + \int_{-\infty}^{t} d\tau_1 \int_{-\infty}^{\tau_1} d\tau_2\; g_4^{(0)}(t-\tau_1, t-\tau_2, x(t))\; i(\tau_1) i(\tau_2) + \\ &+ \int_{-\infty}^{t} d\tau\; g_5(t-\tau, x(t))\; \dot{x}(\tau)\, i(\tau)\end{aligned} \tag{13}$$

—respectively.

Page 30, line 4, please amend the following paragraphs:

~~Where~~ where k_1 is a constant. Since all memory and eddy-current effects have been suppressed in equations (14)-(16), parameter estimation of $L_e(x)$, R_e and k_1 from empirical data will show that they are frequency-range dependent; and, that, furthermore, R_e actually depends upon $x(t)$ since it includes the resistive counterpart to effective coil reactance $L_e(x)$ caused by eddy currents.

Equation (14) is an oversimplification. As recognized in the audio industry, a transducer voice coil is characterized by a frequency-dependent complex effective impedance, which we denote $Z_e(\omega, x)$ to indicate that it also depends upon coil position; it also implicitly depends

upon other, more slowly varying parameters, such as temperature. The effective coil impedance $Z_e(\omega, x)$ characterizes one aspect of the relation between voltage signal $V_{coil}(t)$ applied to the voice-coil circuit on the one hand, and the coil current $i(t)$ caused by this voltage, on the other. This voltage-current relation, or functional, as it is known mathematically, is nonlinear, and furthermore involves electrodynamical memory effects (distributed delays) as described above. In general this relation can be expanded in a functional series of the type known in the literature as a Volterra series. The multivariate coefficient-functions of this Volterra series depend on coil position and motion within the magnetic-circuit airgap.

Current-nonlinear effects, i.e. deviations from linearity of the voltage-current functional, were found to be measureable. For the Labtec Spin70 speaker transducer, one of the large signal data parameters ~~which are~~ illustrated in FIG. 4, namely L_e, was found to vary with $i(t)$ as the coil neared its negative excursion. However, it was also found through modeling, simulation and measurements that current-nonlinear effects in speakers are typically small (at the few percent level), although they can become important for woofers played at high volumes. Thus, for many transducers, the full complexity of the current response $i(t)$ to a given applied voltage $V_{coil}(t)$ can often be usefully approximated by a linear functional relation, in which memory effects (due to eddy currents in the magnetic pole structures, and in the aluminum coil former if any) are still included. This approximate linear relation can be derived from equations (11)-(13) and is expressed as follows:

$$\cancel{V_{coil}(t) = R_e\, i(t) + v(t)\, Bl(x(t)) + \int_{-\infty}^{t} d\tau\ g_3(t-\tau, x(t))\, i(\tau)}$$

$$V_{coil}(t) = R_e\, i(t) + \dot{x}(t)\, Bl(x(t)) + \int_{-\infty}^{t} d\tau\ g_3(t-\tau, x(t))\, i(\tau) \qquad (17)$$

Page 31, line 20, please amend the following paragraph:

The second (velocity dependent) term on the right hand side of equation (17) is the BEMF due to coil motion; the other two terms comprise the EMF due to the overall effective coil impedance. Within the approximation, invoked above, of a slowly changing (low frequency) position $x(t)$, the Fourier transform of g_3 with respect to time is simply the subtracted effective coil impedance in frequency domain, i.e. the coil impedance with the Ohmic coil term subtracted. We denote this subtracted coil impedance as $Z^{sub}{}_e(\omega, x)$. More precisely, when a probe voltage signal at a typical audio (or supersonic) frequency is applied to the voice coil and the attached diaphragm is mechanically held (blocked) at a fixed position x, the effective

impedance, due to the coil's inductance and its interaction with eddy currents and magnetization within the magnetic poles, is by definition- $Z_e(\omega, x) = Z^{sub}{}_e(\omega, x) + R_e$, where the R_e term is added in series and represents the coil's Ohmic resistance (see Equation (17)). Note that the subtracted impedance $Z^{sub}{}_e(\omega, x)$ has both resistive and reactive components; the former is attributable to eddy-current dissipation inside the magnetic poles (and also in the coil former, in case that is made of aluminum). The reactive component of $Z^{sub}{}_e(\omega, x)$ is known in prior art as $L_e(x)$, with the frequency dependence often left implicit, as it was in equations (14)-(15) above.

Page 32, line 18, please amend the following paragraph:

For sufficiently high frequencies, and in the case of non-metallic former, the subtracted impedance $Z_e^{sub}(\omega,\ x)$ arises from currents and EMF's induced in the coil and within a narrow skin layer~~s~~, within the pole structures and adjacent to the coil. For a simple cylindrical geometry with infinite axial extent, $Z^{sub}{}_e(\omega, x)$ is independent of x; in that approximation, Vanderkooy [J. Vanderkooy, J. Audio Eng. Soc., Vol. 37, March 1989, pp.119-128] ~~-~~has shown that the (complex plane) phase angle of the subtracted impedance begins to approach an asymptotic value of 45^o once the frequency increases well above the normal modes of mechanical resonances. Measurements for actual speaker transducers yield a range of possible asymptotic phase angles, both above and below this value [J. D'Appolito: "Testing Loudspeakers", Audio Amateur Press; 1998.] For the ~~LabTec~~Labtec Spin 70 speaker transducer analyzed in the present study, the asymptotic phase angle was measured to be approximately 70^o, varying little with coil/diaphragm position x.

Page 33, line 28, please amend the following paragraphs:

(II) ~~—~~Progressively smaller nonlinear effects can be corrected by applying successive new linearizing filters, and this progression of successive corrections will often *converge* in the sense of perturbation theory.

It should be noted that the ability to systematically apply more and more modular control tiers~~,~~ can be useful even if a higher-tier correction is larger than a lower-tier one.

FIG. 7 is a flow chart ~~which~~ that illustrates the process of linearization in accordance with the present invention. First, a model of a portion of the audio reproduction system is provided in step 1301. Next, a control engineering technique is utilized in the time domain to control an output of the audio transducer system based upon the model, via step 1302.

Page 34, line 20, please amend the following paragraphs:

FIG. 8 is a block diagram of the main portion of a sound reproduction system and a control system for controlling the operation of the sound reproduction system in accordance with the present invention. An audio signal 1401 is input to a controller 1402, which contains algorithms based on a control model, which in turn is based on a physical model (such as the one described by equations (6)-(16) of this section) of the processes within the audio transducer system. These algorithms may be functions of state variables such as acceleration, velocity, and position of the coil/diaphragm assembly. With reference to FIG. 6, the modeled processes may include the signal conditioning process 1102, the voice coil transduction process 1104, and the sound conditioning process 1106, as discussed above. The state variables 1403 from the sound reproduction processes are input to the controller 1402 from a measurement system 1404. The measurement system 1404 consists of a sensor conditioner 1405 and ~~a plurality of~~one or more sensors, 1406a, 1406b, and 1406c, which take measurements of variables from the sound reproduction system. The sensor conditioner 1405 amplifies and converts the signals from the sensors 1406a, 1406b, and 1406c to the state variables 1403, which are provided to the controller 1402. Sensor 1406a may, for example, measure a variable such as current from the drive amplifier 1407. Sensor 1406b may, for example, measure an internal circuit parameter, such as parasitic capacitance, of the transducer 1408. Alternatively, sensor 1406b could electronically measure the impedance of one of the voice coils of transducer 1408, or it could optically measure an indicator of voice coil position. Sensor 1406c may, for example, measure a variable from the acoustic environment, such as sound pressure by using a microphone. By digitizing both the state variables 1403 and the audio signal 1401, and combining them via a DSP, the controller 1402 modifies the audio input 1401, converts it back to an analog voltage, and thus outputs a compensated analog audio signal on line 1409 to the amplifier 1407. The amplifier 1407 outputs a drive signal on line 1410 to the transducer 1408.

The audio transducer state variables ~~which~~ that are measured and fed back to controller 1402 are generalized coordinates of the transducer dynamical system. These generalized coordinates usually vary nonlinearly with the position of the voice coil/diaphragm assembly with respect to the transducer frame, and thus, with suitable calibrations, serve to provide controller 1402 with estimates of recent values of that position. Controller 1402 then uses these real-time position estimates to suitably modify the input audio voltage signal before applying it across the voice coil. Multiple position-indicating signals can be fed to the controller, as depicted in FIG. 8; they are derived from one or more position-indicating generalized coordinates. It may be useful to measure more than one position-indicating generalized coordinates, because in some

portions of the range of coil/diaphragm excursions, it could happen that a given generalized coordinate may not be a monotonic function of coil/diaphragm position, while another generalized coordinate is monotonic in that portion of the range. Thus, the advantage of measuring and feeding back values for multiple generalized coordinates, is that these coordinates may be chosen in such a way that the configuration space of their joint values is approximately a one dimensional differentiable manifold, where the coil/diaphragm position is a continuous and differentiable function on this manifold. And if each of the selected generalized coordinates is also a continuous and differentiable function of coil/diaphragm position, the mapping between a tuple of simultaneously measured generalized coordinates and the corresponding position, is both invertible and differentiable, allowing the use of the tuple to compute the audio signal modification within the controller DSP. One embodiment of this computation, based on a single generalized coordinate ~~which~~ that is derived from infrared optical measurements, is described in detail in Detail 10.

It will be readily apparent to those skilled in the art that additional and different sensors may be utilized, and different signal conditioners may be used to recover state variables and internal parameters from the sensor signals and provide control signals to the system. Additional sensors may include, for example: accelerometers, additional transducer coils, or new coil-circuit elements. Such sensors can provide analog measurements of various voltages appearing in the transconductance equation (14), or of other voltages ~~which~~ that allow the estimation of various terms and state variables in either equation (14) or the mechanical (transduction process) equation (15). State variables and parameters must be identified for each of the sound reproduction processes, and a sufficient set of them must be measured to effect control.

It has been discovered that measurements not usually regarded as state variables can be used effectively in controlling the audio reproduction processes. In the prior art systems, the following variables are typically considered as defining state:

x axial position of coil/diaphragm assembly,

$\dot{x}$ ———axial velocity of coil/diaphragm assembly,

$\ddot{x}$ ———axial acceleration of coil/diaphragm assembly,

i voice-coil current.

What follows is a list of other measurable variables, among them internal parameters characterizing the processes ~~which~~ that are considered constants in small signal analysis, as well as state variables, such as pressure, which would be externally measured (using a microphone in this case). The variables and parameters on this list can all be used in practicing the present invention. Control systems using one or more of these variables and parameters are described

below. Some measurable variables can be measured by reference to other variables through known functional dependencies; for instance, temperature can be inferred from coil resistance and a lookup table. Internal -parameters and other variables not listed in the above list include, for example:

$V(t)$ voice coil voltage,

$i(t)$ voice coil current,

R_e voice coil resistance,

L_e voice coil inductance,

Z_e complex voice coil impedance,

$C_{parasitic}$ voice coil/magnet parasitic capacitance,

$BEMF$ back-EMF,

ϕ complex phase angle of voice coil impedance,

T_e voice coil temperature.

There are other internal parameters such as Bl and K, respectively the motor factor and suspension stiffness. These parameters may be difficult to measure directly, although they can be extracted from measurements of other variables via parameter estimation methods. The voice-coil voltage $V(t)$ and voice coil current~~,~~ $i(t)$ are considered ~~as~~ internal variables, rather than stimuli, because the full audio transduction process according to the present invention includes creating $V(t)$ and $i(t)$ as internal variables.

Page 39, line 13, please amend the following paragraphs:

The control model treats the motor factor $Bl(x)$, the effective coil inductance $L_e(x)$, and the suspension stiffness $K(x)$ as functions of $x(t)$, the current axial position of the coil/diaphragm assembly. These three functions cause most of the nonlinearities, and thus distortions, of audio transducers, as explained above. The motor factor $Bl(x)$ determines the motive force term in equation (19) as well as the BEMF term in equation (18); $L_e(x)$ determines the inductive EMF term in equation (18); while $K(x)$ determines the elasto-acoustic restoring force in equation_(19). In the context of the present invention, these three functions are derived from calibration measurements on the system, which yield the functional dependence of Bl, L_e and K upon x; these functions can, for instance, be obtained from commercially available transducer test equipment such as a Klippel GMBH laser metrology system. In one embodiment

of this invention, the functional dependences $Bl(x)$ and $L_e(x)$ are entirely obtained from such a laser metrology system, while $K(x)$ is obtained by combining knowledge of $Bl(x)$ and $L_e(x)$ with ramped DC-drive calibration runs, as fully described in Details 5 and 10 below.

In transducer operation, the three functions $Bl(x)$, $L_e(x)$ and $K(x)$ must be combined with approximants to a function mapping the measured position-indicator state variable onto the actual position x, as described in Details 4, 5, and 10 below, in order to provide the controller DSP with an estimate for the values of Bl, L_e and $K(x)$ at the ~~current~~ present moment t.

The controller then estimates the BEMF term by multiplying the estimated present value of $Bl(x(t))$ by an estimate for the present velocity $\dot{x}(t)$; the latter may be obtained either from a numerical differentiation of the recent history of discrete position measurements, or from an independent velocity measurement. In one embodiment of the present invention, velocity is estimated via numerical differentiation of estimated position, as described in Detail 10 below. Simulations of the BEMF correction shows that it can be usefully filtered in the frequency domain, as this correction has its greatest effect over a limited frequency range. Such filtering reduces the noise due to the numerical differentiation of position. Once the nonlinear BEMF term $Bl(x)\,\dot{x}$ in equation (18) is thus estimated, it is corrected for by being added by the control circuit to the voltage representing the audio information. A *linear* BEMF term can also be calculated and *subtracted* from the voltage representing the audio information, in order to provide damping if required. The subtracted linear part of the BEMF is chosen such that the effect of the subtraction is to electronically add back a positive constant to the mechanical drag coefficient R_{ms} in equation (19). This positive constant is some adjustable fraction, p, of the Thiele-Small small-signal BEMF contribution to the drag coefficient that would arise due to the equilibrium value $Bl(0)$ *without* any correction. ~~Thus,~~

In many cases of interest the effective coil inductance $L_e(x)$ in equation (18) is very small. If we neglect this inductance, the inductive EMF term $L_e(x)\dfrac{di}{dt}$ in equation (18) disappears, and that differential equation becomes an algebraic equation. With this simplification, the voltage signal that is output from the control circuit to the voice coil in order to compensate for the nonlinear BEMF is:

$$\cancel{V_{coil} = V_{audio} + \left(Bl(x) - p\,Bl(0)^2 / Bl(x)\right)\dot{x} \quad 20)}$$

$$V_{coil} = \frac{Bl(0)}{Bl(x)} V_{audio} + \left(Bl(x) - p\,R_e \frac{Bl(0)}{Bl(x)}\right)\dot{x} \qquad (20)$$

~~Where~~ where V_{audio} is the voltage representing the audio signal before the BEMF correction. Note that other modular corrections may be included in V_{audio}, as described below.

We next turn to the case where the effective coil inductance $L_e(x)$ in equation (18) is not neglected, and ~~describing~~ describe another type of modular control law in the context of the present invention, namely a control law that ~~correcting~~ corrects for the inductive EMF term in equation (18). Like the BEMF control law described above, the inductive control law partially linearizes the transductance sub-process. Specifically, the inductive control law addresses the nonlinearity, and thus distortion, caused by the position dependence of the effective coil inductance $L_e(x)$. In order to derive the inductive control law in as simple a manner as possible, the BEMF term is temporarily ignored in the transconductance equation (18); later in this section, all four of the modular control laws described in the context of this invention (BEMF, inductive, spring and motor factor) will be combined.

Since the embodiment described below for the correction of the inductive EMF term $L_e(x)\dfrac{di}{dt}$ in equation (18) has no history in prior art, the derivation of this correction is presented in some detail here. For simplicity, noise is ignored in this derivation, as ~~is~~ are the deviations of the in-operation digital signal processor (DSP) estimates for $L_e(x(t))$ from the actual values of this variable.

Page 42, line 5, please amend the following paragraphs:

If $V_{audio}(t)$, $L_e(x)$ and $x(t)$ are treated as known functions, equation (23) can be viewed as a linear first-order ordinary differential equation for the unknown function $i(t)$. It is a well-known mathematical fact that this differential equation admits a *unique* solution $i(t)$ for any given causal signal $V_{audio}(t)$, i.e. for an audio input signal that begins at some ~~definite~~ initial time t_0 in the past, given an initial condition $i(t_0) = i_0$. ~~The latter condition can safely be assumed,~~ ~~s~~Since any real-life signal is causal, we can safely assume that there is an initial time t_0 such that $i(t_0) = 0$ and $V_{audio}(t_0) = 0$. ~~On the other hand~~Then, it is easily verified by substitution that a particular solution of the differential equation (23) is given by

$$i(t) = V_{audio}(t) / R_e \qquad (24)$$

The combination of these two facts, namely, that equation (23) has a unique solution for the coil current in terms of the audio voltage input, and that equation (24) is a particular solution of

equation (23) and is valid at an initial time t_0 such that $i(t_0) = 0$ and $V_{audio}(t_0) = 0$,— completes the proof that equation (24) does in fact hold for all values of t. In other words, it has been proven that the coil current $i(t)$ is related to the audio signal $V_{audio}(t)$ by a simple Ohm's law, without any inductive term, provided that BEMF is ignored and that the control law of equation (22) is implemented.

This demonstrates that by simply adding to the audio signal voltage a term that is the derivative of this same audio signal, multiplied by the ratio of the nonlinear inductance to the coil resistance, as done in equation (22), a correction for the effects of inductance alone can be made. In one embodiment of the present invention, the voltage differentiation on the right hand side of equation (22) is implemented numerically by the DSP, as fully described in Detail 10 below; this alone introduces additional terms on the right hand side of equation (24), thus making the elimination of the inductive term approximate, rather than exact-. Furthermore, it will be appreciated from the detailed description of polynomial interpolations in the context of this invention (Detail 10 below) that the correction of the inductive effect by the physical controller, as opposed to the ideal one assumed in the above derivation, is approximate, rather than exact. This caveat would hold even were an exact, analog differentiation to be used by the controller. And it also holds for the numerical BEMF correction described above.

In the case of input to a voice coil ~~which~~ that is used for audio reproduction, removing all the inductance as described in equations (21)-(24) might lead to an equalization problem, since the higher frequencies can be over-compensated. Thus, in one embodiment, an optional linear part of the inductance is added back to endow the audio system with a flatter frequency response. This is described in Detail 10 below.

Page 43, line 32, please amend the following paragraphs:

The correction of the nonlinear electromechanical effects in the mechanical (transduction) equation of motion (19) is based upon a derivation similar to, but different from, the standard control theory derivation of a control equation presented in the Background section above as prior art. One practical problem with the mechanical equation (19) as a starting point for a control model~~,~~ is that the inertia term involves the coil/diaphragm acceleration $\ddot{x}$. This term increases rapidly with frequency, eventually becoming too large to be considered in a compensation system. However, because the acoustical radiation efficiency of the cone also increases with frequency, the inertia non-compensation is balanced by the radiating efficiency, within limits. This trade-off is known in prior art to result in a more or less constant output over a range of frequencies referred to as the 'mass controlled' range. Transducers are normally

designed with this effect in mind.

By ignoring mass in equation (19), that is to say by neglecting inertial effects, the following first order differential equation is obtained:

$$R_{ms}\,\dot{x} + xK(x) = Bl(x)\,i(t) \qquad (25)$$

In the general nonlinear state space form, equation (25) is recast thus:

$$\cancel{\dot{x} = \phi(x) + \psi(x)\,u(t)}\ \dot{x} = \phi_1(x) + \psi_1(x)\,u(t) \qquad (26)$$

—where,

$$\cancel{\phi(x) = \frac{-xK(x)}{R_{ms}}}\ \phi_1(x) = \frac{-xK(x)}{R_{ms}}\ ,\cancel{\psi(x) = \frac{Bl(x)}{R_{ms}}}\ \psi_1(x) = \frac{Bl(x)}{R_{ms}}$$

(27)

—and:

$$u(t) = i(t) \qquad (28)$$

Following the feedback linearization approach, consecutive derivatives of the transducer output are taken until its input, $u(t)$, appears in one of the derivatives. But that is already the case in equation (26), which, when combined with equation (27), yields for the first derivative of coil/diaphragm position $x(t)$:

$$\dot{x} = \frac{-xK(x) + Bl(x)u(t)}{R_{ms}} \qquad (29)$$

Note that the input, $u(t)$, indeed appears explicitly in the first derivative of the position state variable, x.

The controller linearizing the transduction process should cause the transducer output $\dot{x}(t)$ to be proportional to the audio input. Equating $\dot{x}(t)$ with $V_{audio}(t)$ in equation (26) and solving for $u(t)$, and assuming that the function $\cancel{\psi(x)}\ \psi_1(x)$ defined in (27) is nonsingular, we obtain:

$$\cancel{u(t) = [\psi(x)]^{-1}[-\phi(x) + w]}\ u(t) = [\psi_1(x)]^{-1}[-\phi_1(x) + w]$$

(30)

—where $w(t)$ is the generator or reference (in our case the audio program input $V_{audio}(t)$ to the uncorrected transducer), and $R_e\,u(t)$ is the actual voltage input to the voice coil in the controlled (corrected) transducer if the signal conditioning process is ignored. Substituting and rearranging terms in equations (27), (28) and (30), provides:

$$i(t) = \frac{xK(x)}{Bl(x)} + w\frac{R_{ms}}{Bl(x)} \qquad (31)$$

By applying this (ideal) control equation to the second order differential transduction equation (19), it is possible to see whether the latter is thereby linearized.

Substituting equation (31) into equation (19) provides:

$$\cancel{m\ddot{x} + R_{MS}\dot{x} + K(x)x = Bl(x)\left[\frac{xK(x)}{Bl(x)} + w\frac{R_{ms}}{Bl(x)}\right]}$$

$$m\ddot{x} + R_{ms}\dot{x} + K(x)x = Bl(x)\left[\frac{xK(x)}{Bl(x)} + w\frac{R_{ms}}{Bl(x)}\right] \qquad (32)$$

This leaves,

$$m\ddot{x} + R_{ms}\dot{x} = wR_{ms} \qquad (33)$$

Equation (33) is a linear differential equation with constant coefficients. Note that from the above a general method of linearizing this form of nonlinear dynamical equation is presented, and any further linear terms can be added to the equation without changing the validity of the linearization approach.

Lumping the terms of the rearranged control equation (31) and using equation (28) provides the following form of the transduction control equation:

$$u(t) = S(x) + w(t)\,B(x) \qquad (34)$$

~~Where~~ where $S(x)$ and $B(x)$ are functions of position and $w(t)$ is the audio information.

Equation (34) provides a correction for the open loop non-linear transfer function of the speaker transducer, provided that the dependencies of $S(x)$ and $B(x)$ on x are known and that real-time measurements or estimates of x are made available to the controller during transducer operation.

Page 46, line 14, please amend the following paragraph:

Clearly, the control law given by equation (34) removes all restoring force due to the spring; a thus corrected transducer would not be stable. Thus a linear (non-distorting) restoring force must be subtracted from $x\,K(x)$. The magnitude of the effective spring constant of this residual electronic linear restoring force~~,~~ can be selected based on the required resonant frequency. This then in effect reduces the transducer operation to the linear case of zero motor factor and a linear (Hooke's law) elastic restoring force. A full description as to how this subtraction is implemented in one embodiment of the present invention, is presented in Details 5 and 10 below.

Page 46, line 32, please amend the following paragraphs:

The control model of equation (34) applies only to the transduction process itself; i.e. it is based on a model of the current to velocity transduction process, and does not cover the process

of injection of current into the coil (the signal conditioning process); nor does it cover the radiation of the sound waves out of the speaker enclosure into the acoustic environment (the sound conditioning process). Likewise, the control models of equations (20) and (22) above, suitably combined, eliminate or reduce only those nonlinearities arising from the transconductance component of the signal conditioning process, but do not correct either of the other two processes (transduction or sound conditioning). And all of the above control laws can, and have been, applied together, or in various partial combinations, in the context of the present invention. This illustrates the modularity of the control approach described as part of the present invention, as discussed in Detail 1 above. Furthermore, the transduction control law of equation (34) can be subdivided into "spring correction" and "motor factor" modular units; e.g. if only the first term on the right hand side of equation (34) is used, this represents a control law which only linearizes the elastic restoring force. Thus, the number of modular control laws described by the above equations can actually be counted as four: BEMF, inductive, spring, and motor factor.

If a choice is made to simultaneously implement all of these modular corrections: the BEMF correction (equation (20)), the inductive correction (equation (22)), and the transduction corrections (equation (34)), this can for example be done as follows. The last term ~~BEMF correction~~ of equation (20) is added to the voltage given by the right-hand side of equation (34); ~~and~~ then the new overall voltage, $u_1(t)$, still in the digital domain, is numerically differentiated (as described in Detail 10 below), and this numerical derivative is ~~finally~~ combined with $u_1(t)$ itself in accordance with equation (22). Finally, the BEMF correction term of equation (2) is added to the new voltage. The overall combined control model for the coil voltage is thus as follows:

$$\cancel{u_1(t) = S(x) + wB(x) + (Bl(x) - pBl(0)^2 / Bl(x))\,\dot{x}(t)}$$

$$u_1(t) = S(x) + wB(x) - pR_e\frac{Bl(0)}{Bl(x)}\dot{x}(t) \ , \qquad (35)$$

$$\cancel{u(t) = u_1(t) + \frac{L_e(x)}{R_e}\dot{u}_1(t)}\ u_V(t) = u_1(t) + \frac{L_e(x)}{R_e}\dot{u}_1(t) + Bl(x)\dot{x}\ . \qquad (36)$$

where

$$V_{coil}(t) = u_V(t)\ . \qquad (37)$$

As explained above, the precise order in which the modular corrections are applied is not very important, as has in fact been demonstrated in the context of this invention.

In order to add back an effective electronic linear restoring force, as discussed above and in Detail 5, the term $S(x)$ on the right-hand side of equation (35) must be replaced by the subtracted version,

$$S(x) - \frac{q\,xK(0)}{Bl\,(x)}$$ (~~36a~~38)

where q is the fraction of the uncorrected suspension stiffness at equilibrium that is added back electronically. Thus equation (35) now becomes,

$$\cancel{u_1(t) = S(x) - \frac{q\,xK(0)}{Bl\,(x)} + w\,B(x) + (Bl(x) - p\,Bl(0)^2 / Bl(x))\,\dot{x}(t)\ ,}$$

~~(37)~~ $$u_1(t) = S(x) - \frac{q\,xK(0)}{Bl\,(x)} + w\,B(x) - p\,R_e\,\frac{Bl(0)}{Bl(x)}\,\dot{x}(t)\ ,$$ (39)

while equation (36) remains unchanged.

In case a choice is made to implement only the transduction correction law, it is still necessary to perform the suspension stiffness subtraction, for stability purposes – as explained above. Thus, the full transduction control law in accordance with the present invention is the following modified version of equation (34):

$$u(t) = S(x) - \frac{q\,xK(0)}{Bl\,(x)} + w\,B(x)$$ (~~38~~40)

One view of the control method described in this invention is that it belongs to the genre of feedback linearization controllers. The transconductance component of the signal conditioning process, and the transduction process, together may be thought of as a dynamic system with voltage input and displacement output. The dynamics of this system are governed by a physical model that can be represented as a three-state system with current, displacement, and velocity as its state variables. As seen above, despite the interactions among all processes comprising the audio reproduction system, various processes and sub-processes can be separately controlled according to this invention by applying only one of the separate basic linearization control laws encoded by equations (20),(22), and (34), or these control laws may be applied in various combinations – depending on user preferences. One option is to apply all of them, as encoded in equations (36) and (~~37~~39), as well as in equations ~~(56)-(59)~~(61)-(64) in Detail 10 below.

Page 49, line 14, please amend the following paragraphs:

FIG. 9 shows the feedback linearization process 20400 with the control law of equation (34), which only linearizes the transduction component of the signal conditioning process, without an electronically restored linear restoring force. The audio signal, $V_{audio} = w$ 20401, is

input to a Linear Compensation Process module 20402 (henceforth abbreviated as LCP). The LCP 20402 multiplies w by the compensation function $B(z)$, where z 20411 is the estimated present value of the position variable. The present value of position variable z 20411 is obtained from the transduction module 20408 of the three-state overall transducer system, via a two step process: first the position indicator state variable $f(x)$ 20413 is measured by the positional sensor module 20412, and then the value of $f(x)$ 20413 is fed as input to a sensor inversion module 20414, which estimates actual position x via an interpolation method as described in Details 5 and 10. Actual position x 20409 and actual velocity $\dot{x}$ 20410 are fed from the output of transduction module 20408 back into the input of the transconductance module 20406, via the physical system itself (not as measured data). The estimated x value, z 20411, is ~~inputted~~ fed into to the LCP 20402 and also to an S-lookup module 20415. The output of module 20415, $S(z) \approx S(x)$ 20416, as well as the LCP output $B(z)w$ 20403, are both fed as inputs to a ~~summer~~summing junction 20404, the output 20405 of which is the corrected audio signal (V_{coil} of equation (34)). This corrected audio signal 20405 is provided as input to the transconductance module 20406 of the three-state transducer system. The current output I_{coil} 20407 of the transconductance module 20406 is provided as input to the transduction module 20408.

FIG. 10 shows the feedback linearization process 20500 for the control law given by equation (~~38~~40); again only transduction corrections are made, but now a linear spring constant (suspension stiffness) is electronically added, as explained above and in Detail 5. The audio signal, $V_{audio} = w$ - 20501, is input to an LCP module 20502. The LCP 20502 multiplies w by the compensation function $B(z)$, where z 20514 is the estimated current value of the position variable. Value z 20514 is obtained from the transduction module 20508 of the three-state overall transducer system, via a two step process as in FIG. ~~204~~9: the positional sensor module 20511 outputs the measured position indicator state variable $f(x)$ 20512, and measured state variable $f(x)$ 20512 is fed as input to a sensor inversion module 20513, which estimates actual position x via the interpolation method. Actual position x 20510 and velocity $\dot{x}$ 20509 are fed back from the output of the transduction module 20508 to the input of the transconductance module 20506 via the physical system itself.

The estimated x value, z 20514, is this time ~~inputted to~~fed into three modules: to the LCP 20502, to an S-lookup module 20516, and to a new 'Electronically Restored Linear Spring'(henceforth ERLS) module 20517. The output of module 20516, $S(z) \approx S(x)$ 20415, as well as the LCP output $B(z)w$ 20503 and the output 20518 of the ERLS 20517, are all fed as inputs to a ~~summer~~summing junction 20504, the output 20505 of which is the corrected audio

signal (V_{coil} of equation (34)). The corrected audio signal 20505 is provided as input to the transconductance module 20506 of the three-state transducer system via the physical system.

FIG. 11 shows the feedback linearization process 20600 for the control law given by equation (~~37~~39) alone, without the inductive correction (36); i.e. for a control law correcting for spring, motor factor and BEMF nonlinearities, including an electronically restored linear spring and electronically restored contribution to the linear drag force term, as explained above. The audio signal, $V_{audio} = w$ 20601, is input to an LCP module 20602. The LCP 20602 multiplies w by the compensation function $B(z)$, where z 20622 is the estimated present value of the position variable. The output $B(z)w$ 20603 of the LCP module 20602 is provided as input to the ~~summer~~summing junction 20604. Value z 20622 is obtained from the transduction module 20610 of the three-state overall transducer system, via a two step process as in the previous figures: the *positional sensor module* 20613 outputs the measured position indicator state variable $f(x)$ 20614, which is then fed as input to a *sensor inversion module* 20615. Sensor inversion module 20615 estimates *actual* position x via the interpolation method. And as in previous figures, the actual position x 20612 and velocity $\dot{x}$ 20611 are fed back by the actual physical system from the output of the transduction module 20610 to the input of the ~~transduction~~ transconductance module 20608. The estimated x value, z 20622, is now ~~inputted to~~fed into *four* modules: to the LCP 20602; to the S-lookup module 20618; to an ERLS module 20620; and finally, to a BEMF-computation module 20616, which applies a numerical differentiation operation D to z 20622. The output 20619 of the module 20618, as well as output 20621 of module 20620 and output 20603 of the LCP 20602, are summed in the ~~summer~~summing junction 20604. The output 20605 of ~~summer~~summing junction 20604, along with the output 20617 of the BEMF-computation module 20616, are provided as inputs to a second ~~summer~~summing junction 20606; finally, the output 20607 of the second ~~summer~~summing junction 20606 is the corrected V_{coil}, which is provided as analog input to the transconductance module 20608 of the three-state transducer system. And the analog coil current I_{coil} 20609, output by the transconductance module 20608, is provided by the physical transducer as input to the transduction module 20610.

FIG. 12 shows the feedback linearization process 20900 for the control law given by equations (36) and (~~37~~39), i.e. implementing all the corrections described in this section, and also implementing two numerical Low Pass Filters: one between the position-indicator variable measurement and the sensor inversion, and another after the computation of the fully corrected coil voltage and before it is fed as input to the coil. The audio signal, $V_{audio} = w$ 20901, is input to

an LCP module 20902. The LCP module 20902 multiplies w by the compensation function $B(z_f)$, where z_f 20921 is a filtered version of the estimated present value of the position variable. The output ~~$B(z)w$~~ $B(z_f)w$ 20903 of the LCP module 20902 is provided as input to the ~~summer~~summing junction 20604. Value z_f 20921 is obtained from the transduction module 20910 of the three-state overall transducer system, via a *three* step process: the *positional sensor module* 20912 outputs the measured position indicator state variable $f(x)$ 20913, which is then fed as input to the low pass filter LPF2 20924, the role of which is to suppress sensor noise; LPF2 would typically roll off at $1-2\ kHz$. The output 20925 of LPF2 20924 is fed to the *sensor inversion module* 20914. Sensor inversion module 20914 again estimates *actual* position x via the interpolation method, in the digital domain; while the actual position x 20911 and velocity $\dot{x}$ 20912 are fed via the physical transducer plant, back from the transduction module 20910 to the transconductance module 20908. The estimated x value, now called z_f 20921,- is ~~inputted to~~fed into the following three modules: to the LCP 20902, to the ERLS module 20920, and to the BEMF-computation module 20915. The S-lookup module 20917 receives its input this time from the filtered, but *not* inverted, positional indicator variable measurement result 20925. The outputs of the four modules 20915, 20917, 20919 and the LCP 20902, labelled respectively 20916, 20918, 20920 and 20903, are summed in the ~~summer~~summing junction 20904. The output 20905 of ~~summer~~summing junction 20904 is passed to an inductive-correction module 20927, which again applies a numerical differentiation operation D, this time to the numerical output voltage 20926 of the ~~summer~~summing junction 20904. The output 20906 of the inductive-correction module 20927 is provided along with numerical output voltage 20926 multiplied by R_e to a second ~~summer~~summing junction 20928, whose output- 20907 is fed to the low pass filter LPF1 20922. The low pass filter LPF1 20922 implements a (partial) correction for the voice coil inductance at equilibrium. The output 20923 of LPF1 20922 is finally fed as the corrected analog voltage V_{coil} to the transconductance module 20908 of the three-state transducer system. As in the previous figures, the physical transducer plant provides the analog output current I_{coil} 20909, output by the transconductance module 20908, as input to the transduction module 20910.

As emphasized above, the present invention requires at least one state variable to be measured in operation for any given run. In the control diagrams depicted in FIG. 9, FIG. 10, FIG. 11 and FIG. 12, it has been assumed for convenience that only a single state variable is measured (although at least two variables, such as for example the position measured by an IR

sensor x_{ir} and the position measured by a laser sensor $x_{lsr} \approx x$, would need to be measured during offline calibration runs in order to derive an interpolated function $f(x)$).

The process of applying a state variable feedback law based on a plurality of measurements of one or many state variables~~,~~ is depicted in FIG. 13. The process 21000 begins with one or several measurements of a state variable or variables from a plurality of sensors, 21001 through 21002. For example a transducer's coil/diaphragm displacement, x, may be measured both via the parasitic capacitance method (Details 7 and 12 below) and the IR method (Details 8 and 13 below). The respective state variable measurement signals, 21003 through 21004, are passed from the sensors to the state estimation module 21005, which synthesizes the desired partial or full state variable estimate, 21006, which in general is a vector state variable. This state variable estimate 21006 is in turn used in the application of the control law 21007 in place of the actual state variable.

For all practical purposes, none of the sensors, 21001 through 21002, can measure its intended state variable exactly. The measurement is always corrupted to some extent by factors including nonlinearities in the measurement, measurement noise, quantization noise, systematic errors, etc. The task of the state estimation module 21005 is to mitigate these corrupting effects. This task may include all or some of the following ingredients: inverting the nonlinearities of the sensors to provide a more linear response to the measurements 21001 through 21002; adaptation to minimize the sensitivity of the state variable estimate 21006 to parametric uncertainties in the measurement, such as uncertainty in gain; filtering the measurement signals 21003 through 21004 to minimize the effects of noise; or fusing multiple measurements of a state variable into one state variable estimate 21006. In addition, many engineering objectives are taken into consideration in the design of the state estimation module 21005. The tradeoffs include such ~~desiderata~~ desirable properties as simplicity of-design, overall reduction in the effects of noise in the system, minimization of the order of the state variable estimator, and cost of implementation. For example, one possible method by which to invert the nonlinearities in any of the measurements 21001 to 21002~~,~~ is via a lookup table based upon offline calibration runs; another possible method, also based upon offline calibration, is via polynomial expansion. The latter is the method used in one embodiment of the present invention, as described in Detail 10 below. Noise reduction may be accomplished by filtering, for example by using finite impulse response (FIR) or infinite impulse response (IIR) digital filters, or else analog filters. The structure of an IIR noise reduction and data fusion filter, and its coefficient values may be determined by trial and error or by analysis. For example, a positional estimation filter could be designed via Kalman filtering techniques, in which a stochastic model of the input signal and state

measurement noise is combined with a model of the transconductance and transduction dynamics (such as equations (18)-(19) above) to resolve the order and coefficient values of the estimation filter. One skilled in the art will realize that various different filtering techniques can be used.

Page 55, line 19, please amend the following paragraphs:

As shown above, the nonlinearities in the electromechanical equations (18) and (19), which result from the position dependence of the $L_e(x)$, $K(x)$ and $Bl(x)$, produce a nonlinear response in the transduction output x as a functional of the voltage input $V_{coil}(t)$. In-operation measurement of at least one position-indicator variable, together with suitable DSP computations as described above and in Detail 5 below, is used to calculate approximations to $x(t)$, $\dot{x}(t)$, $L_e(x(t))$, $K(x(t))$ and $Bl(x(t))$ at any given moment during transducer operation. These numbers, together with the audio program input $V_{audio}(t)$, are then used by the controller circuit to implement a nonlinear feedback law for the transducer voltage input, $V_{coil}(t)$, based on the physical model of the system, as described by the control models given in equations (20), (22) and ~~(38)~~(40). The overall control model obtained by combining the three control laws given by equations (20), (22) and ~~(38)~~(40), namely that given by equations (36) and ~~(37)~~(39) above, was implemented in one embodiment of the present invention; the measured power spectrum distribution for a standard two-tone test, both with the combined correction and with no correction at all, are presented for this embodiment in Detail 14 below. It is seen that the effect of this combined feedback law is to eliminate or greatly reduce the distortions of the 3" Audax speaker transducer for which the data of Detail 14 were taken. Both intermodulation and harmonics peaks were significantly reduced.

In the course of the derivation of the control laws in this section, it was noted that the physical audio transducer parameters $L_e(x)$, $-K(x)$ and $Bl(x)$, as well as the position state variable x, are not perfectly known, and that for that reason, full correction as it appears in the equation of this section, will not in fact occur. The equations were derived assuming perfect knowledge by the controller; this was done to make the derivation of the control laws more transparent. In practice, however, these physical parameters and state variables are close estimates of their actual values. The attendant errors in modeling and measurement – both systematic and noise errors - introduce a small amount of ~~unmodeled dynamics~~uncertainty in the system.

It is a well--known result in control theory that under certain conditions, unmodeled dynamics can lead to instabilities in a dynamical system under feedback. Care has been taken in

the implementation of the feedback laws of this section to reduce the sensitivity of the electromechanical system to ~~these unmodeled dynamics~~this uncertainty, thus preventing the possibility of dynamic instability in the electromechanical system, provided the coil/diaphragm excursion is not too high.

Anyone skilled in the art will realize that other processes and process-components can be included in the transducer physical model, in addition to the transconductance and transduction ~~which~~ that are respectively encoded in the electric and mechanical equations (18)-(19). Examples of such additional processes are frequency partitioning and sound conditioning. These can be included in both the physical and control models, in accord with the modular approach to control modeling and implementation described in Detail 1 above. Similarly, the control models herein described can also be improved by accounting for ~~other~~smaller effects and terms within the electromechanical physical model, such as the terms that are not present in equations (18)-(19) but are present in equations (6) through (16).

DETAILED DESCRIPTION 3: JUSTIFICATION OF APPROXIMATIONS

A simplified physical model of a general speaker transducer, together with a modular collection of control models designed to implement linearization filters for sub-processes within the physical model, were presented in Detail 2 above. There are two ways in which these mathematical models are used in the context of the present invention: in actual physical implementation, and in simulation.

Page 58, line 1, please amend the following paragraphs:

As explained in Detail 1 above, there are many nonlinearities in the physical processes governing transducer operation, such as nonlinear elastic restoring force (i.e. nonlinear effective spring "constant"); nonlinear motor factor; nonlinear effective voice coil inductance; and motor BEMF, to name the most important ones. Computer simulations based upon the transducer-plus-controller model (and thus incorporating the leading nonlinear processes listed above) were used in the present work to study the effect of all of these nonlinearities, thereby elucidating the merits of implementing partial correction for a subset of the nonlinearities. For instance, it was found via simulation that transconductance nonlinearities (BEMF and inductive) are responsible for significant audio distortions at various important frequency ranges, which- led to the inclusion of corrections for these effects in -the control law (equations (20) and (22) above). In fact, dependent on program material, correcting for non-linear spring effects can have the consequence of increasing the excursions of the transducer coil/diaphragm assembly and thus

increase the non-linear effects of BEMF and $L_e(x)$. Nevertheless, it is still possible to achieve improved audio performance, especially at he low end of the audio spectrum, by correcting only for the nonlinearities in effective spring stiffness and in the motor factor. This~~,~~ fact, as well, had been predicted by simulations of the model, and corroborated by experiment.

We present several ~~few~~ key simulation results relevant to the invention herein disclosed.

FIG. 14 shows curves 4100 of simulated Power Spectral Density (PSD) which illustrate the effect of the transduction corrections alone (spring stiffness and motor factor correction, equation (34)) both with and without BEMF and nonlinear inductance in the system. In FIG. 14 the vertical axis is a measure of PSD in relative dB units. The curves of FIG. 14 were generated by simulating the performance of a particular transducer (that of the Labtec Spin 70 speaker) using a single $100\,Hz$ tone; each curve clearly ~~shown~~ shows that the highest power is in the fundamental $100\,Hz$ tone, but that significant power is also present in the various harmonics of this tone. Overall, the curves of FIG. 14 shows that even at frequencies where BEMF is significant, introduction of corrections for spring and force constant greatly improve the system performance. Curve 4103 depicts the simulated PSD with no BEMF voltage term modelled, with *linear* (i.e. position independent) inductive EMF voltage term modelled, and with no correction incorporated in the modelling; the harmonics, and power present at non-harmonic frequencies, are an artifact of the finite time windowing used to perform the FFT (Fast Fourier Transform) in the simulation. Curve 4101 shows the PSD when the position-dependent (nonlinear) BEMF and position-dependent inductive EMF voltage terms both modelled, but still with no correction; the harmonics, as well as the general diffuse high-frequency content of the power spectrum, are seen to be enhanced by nonlinearity-caused distortions. Curve 4102, again depicting the PSD with nonlinear BEMF and nonlinear inductive EMF, but this time *with* transduction corrections, shows a marked decrease in harmonics and other, diffuse high-power spectral content. Finally, curve 4104 depicts the PSD with no BEMF and with linear inductive EMF, as in curve 4103, but with the difference that the transduction correction is applied.

It is inevitable that there will be some delay between measuring and reading the sensor output, and sending out the command to compensate for the position-dependent nonlinear spring stiffness and motor factor (and for any other nonlinearities for which terms are included in the controller). Using model-based simulation, it was possible to determine that the existence of this delay, while somewhat degrading- the performance of the control algorithm, did not cause a significant problem, nor did it render the algorithm ineffective.

The curves of FIG. 15 illustrate the reduction in distortion as a function of the delay in the correction loop 4200. As in FIG. 14, the vertical axis is a measure of relative PSD

magnitudes in dB. The curves of FIG. 15 depict the simulated PSD of the transducer-cone velocity, again for a $100\,Hz$ audio input tone. In obtaining these simulation results, it was important to keep the amount of the nonlinearities the same for all the cases that were considered. This was achieved by suitably scaling the driving force as the time delay was varied. It is clear, from the curves of FIG. 15, that longer delays in the correction loop will increase distortion. However, for a $100\,Hz$ tone, even at $200\,\mu\sec$ delay, the distortion is seen to be less than that of the uncorrected system. Curve 4201 depicts the PSD with no correction; curve 4202 depicts the PSD *with~~(~~ ~~transduction)~~ correction* but for the ideal case of no delay; while curves 4203 and 4204 show the PSD curves with correction modelled and with simulated delays in the amounts of $100\,\mu\sec$ and $200\,\mu\sec$, respectively.

While a complete nonlinear spring cancellation will reduce the distortion in a speaker's acoustic output, it will also remove the restoring force that was provided by the mechanical spring in the uncorrected speaker transducer, as discussed in Detail 2 above. In order to keep the speaker cone centered near its equilibrium position and place the mechanical resonance of the speaker at the desirable frequency, linear stiffness can be added electronically, as seen in Detail 2 above. FIG. 16 displays a plot 4300 depicting the position of cone (i.e. the axial position of the coil/diaphragm assembly) in the presence of a single-tone excitation. Without the added electronic contribution to the effective spring stiffness, the cone may drift from its equilibrium position, and may reach its limit of excursion; this is illustrated in the simulation shown in curve 4302. Curve 4301 shows the corresponding simulated time-dependent cone excursion when an electronically-_added linear spring constant (suspension stiffness) is incorporated in the model.

Page 61, line 18, please amend the following paragraph:

FIG. 19 is a simulated version of spectral plot results 4600 of the two-tone intermodulation and harmonic distortion test for which actual, physical implementation results are reported in Detail 14 below. The two input tones are at $60\,Hz$ and $3\,kHz$, and the portion of the simulated power spectrum distributions (PSDs) shown in the curves of FIG. 19 are in the vicinity of ~~the~~ $3\,kHz$. The curves (4601 through 4603) clearly show the forest of intermodulation peaks, spaced uniformly $60\,Hz$ apart and with decreasing power level away from the $3\,kHz$ main peak. As is the case for the real spectrum in this frequency region (FIG. 65), the simulation shows the intermodulation peaks to be significantly suppressed when all four linearizing-filter corrections are applied (i.e. with the combined correction law given by equations (36)-~~(37)~~(39)). But unlike in the physical implementation, it is possible to select arbitrary time delays in the simulation. Two different delay values were chosen for this simulation: $10\,\mu\sec$ and $50\,\mu\sec$.

And delays were only applied for the corrected runs. Curve 4601 shows the simulated uncorrected PSD; curve 4602 shows the dramatic intermodulation reduction when the corrections are applied, with $10\,\mu\mathrm{sec}$ simulated delay. Finally, curve 4603 shows the simulated PSD with corrections and with the longer simulated delay of $50\,\mu\mathrm{sec}$.

Page 62, line 10, please amend the following paragraphs:

It will be clear to those skilled in the art that simulation of any particular implementation of the linearization and control methods described in this disclosure provides valuable information for practically implementing such systems for any particular application; and, furthermore, that the simulations developed here can be greatly expanded to cover many such systems and applications.

DETAILED DESCRIPTION 4: STATE MEASUREMENT THEORY

The present invention is described in the context of controlling an audio reproduction system, in part, by a model requiring real time measurement of at least one position-dependent state variable of the speaker transducer. In particular, one such state variable is the axial position x of the coil/diaphragm assembly. Real-time values of the state variable x are needed during transducer operation in order to effect the linearization of the transconductance and transduction processes, as set out in Detail 2. According to the present invention, it is unnecessary to have a direct measurement of x; it suffices to measure, instead, a position-indicator state variable, i.e. a variable which varies monotonically (but, in general, nonlinearly) with x within the range of possible diaphragm excursions. Once this position-indicator nonlinear state variable $f(x)$ is calibrated against x, real time measurements of the state variable $f(x)$ can be used by the controller to effect linearization.

Page 63, line 16, please amend the following paragraphs:

The IR method is fully described in Details 8 and 13. The Z_e method is fully described in Details 6 and 11. The C method is fully described in Details 7 and 12. The position information derived by Z_e and C methods is generated using internal electronic parameters of the transducer. In contrast, the IR method is based on an external measurement of position. In all cases, to be useful as stand-~~-~~alone position indicators, the respective variables must be monotonic, but not necessarily linear, with position. It will be appreciated that there are other possible position indicators according to the present invention, which are measurable from internal electronic circuit parameters of the transducer that are not constant during transducer

operation, but instead vary monotonically with x. One of ordinary skill in the art will readily recognize that there are many measurements that can be made on an audio transducer, but that $K(x)$, $Bl(x)$, and $L_e(x)$ are commonly presented as the parameters most responsible for the nonlinearities in the operation of such a transducer. The relationship of these parameters to these nonlinearities was explained in detail in previous sections, as was the fact that $L_e(x)$ also varies somewhat with frequency and depends on temperatures in the coil and within the magnet assembly.

As an example of the use of position-indicator measurements in the controller in the context of the present invention, we consider one of the sub-process linearization laws presented in Detail 2 above; namely, the transduction-process control equation (34), where the transduction parameters S and B are non-constant functions of x. Any nonlinear position-indicator state variable $f(x)$ can be substituted for x, as long as the positional related information is monotonic with x and is well behaved over the range of interest, i.e. the range of coil/diaphragm excursions in actual audio operation over which the correction is required. In other words, a nonlinear expansion in x can be replaced by a nonlinear expansion in any measurable variable that has a monotonic relationship with x over a suitable range of values. Thus, the variables S and B can be redefined as functions of x_{ir}, L_e, Z_e or $C_{parasitic}$, depending on the positional-detection method selected. The control law (34) then assumes the following different forms:

$$\cancel{i(t) = S(x_{ir}) + wB(x_{ir}) \quad (45)} \; \underline{i(t) = S_{ir}(x_{ir}) + wB_{ir}(x_{ir})} \qquad \underline{(41)}$$

$$\cancel{i(t) = S(L_e) + wB(L_e) \quad (46)}$$

$$\cancel{i(t) = S(Z_e) + wB(Z_e) \quad (47)}$$

$$\cancel{i(t) = S(C_{parasitic}) + wB(C_{parasitic}) \quad (48)} \; \underline{i(t) = S_L(L_e) + wB_L(L_e)} \qquad \underline{(41a)}$$

$$\underline{i(t) = S_Z(Z_e) + wB_Z(Z_e)} \qquad \underline{(42)}$$

$$\underline{i(t) = S_C(C_{parasitic}) + wB_C(C_{parasitic})} \qquad \underline{(42a)}$$

Page 65, line 11, please amend the following paragraphs:

In accordance with the present invention, the transduction control law, equation (34), has been used to illustrate the use of nonlinear position indicators for linearization corrections. However, the same indicators can be used for some of the other corrections that can be added in a modular fashion to any particular implementation. These combinations of the modular control laws, described in the context of the present invention, are given by the control equations (20), (22), and (36)-~~(37)~~(39) in Detail 2 above. In the case of the BEMF correction (equation (20)), the motor factor $Bl(x)$ can be stored in the controller as a function of the nonlinear state variable $f(x)$, while the instantaneous velocity $\dot{x}$ can be obtained not by measuring a motional state variable, but rather via numerical differentiation of the position, which in turn is obtained from $f(x)$ via the stored inverse functional relation f^{-1}. All controller-stored functions, whether having the form of polynomials, look-up tables or splines, or some combination of the these, will be computed, based upon calibration or characterization of the transducer, 'offline'; i.e. before actual transducer operation.

Page 66, line 1, please amend the following paragraphs:

Information from other external measurement apparata not utilized in the context of this invention, such as accelerometers, microphones, voltages from additional coils and/or additional transducers, can also be used to provide additional state variables, and thus can be used to add precision to, or reduce the noise, for positional or motional estimates.

DETAILED DESCRIPTION 5: S AND B MEASUREMENT THEORY

The present invention is described in the context of extracting the positional state of the speaker transducer's coil/diaphragm assembly, in operation, using measured state variables, from either internal circuit parameters, or signal(s) from external position-sensitive device(s), that are variables with that position. Measurement of all the parameters required to estimate S and B (the transduction-process variables introduced in Detail 2 above) with commercially available test equipment is both time consuming and fruitless. For a viable control scheme, the parameters must be regularly updated as they are sensitive to both time and temperature changes.

Accordingly, a method to measure S and B in a timely manner is described. The method used in this embodiment of the invention, and described in this section, to make the current value of B available to the controller DSP during operation, is also utilized for the electrodynamical transducer parameters Bl and L_e, as described in Detail 10 below. The values of Bl and L_e are

needed by the controller in order to implement the transconductance corrections, namely the BEMF and inductance corrections respectively, as explained in Detail 2 above.

Page 67, line 5, please amend the following paragraphs:

Values for S can be measured directly from the control loop 6100. Considering the linearization correction equation (34) (or its subtracted version, equation ~~(38)~~(40)) for the transduction process alone with no audio information w, and hence without the B term, the spring force term S can be output independently simply by outputting a DC value – because for a DC signal, the only force in the correction equation is the static (spring-force to motor-factor ratio) term $S(x)$, and the numerical value of S can thus be measured. And since the corresponding numerical DC value of the arbitrary measure of position $f(x)$ is also measured and fed back to the controller 6101, the approximate functional dependence of S upon f can be extracted via a suitable polynomial fit, and then used by the digital controller 6101 to look up the value of S which goes into real-time linearization correction of an actual, AC audio signal.

FIG 21 is a flow diagram of a process for determining S as a function of position of a transducer. FIG 22 shows the voltage waveform 6206, the current from which is utilized to move the cone of transducer 6103 and thus to determine and plot S as a function of x. Waveform 6206 is output in step 6201 ~~which~~ that moves the diaphragm through positive and negative values of position x, relative to the no-drive equilibrium value $x = 0$, over the range of the transducer's excursion. If, as is the case in the current embodiment~~,~~, a voltage-~~-~~controlled amplifier ~~-~~is used, a voltage ramp 6206 is output from controller 6101 as shown. After a new discrete voltage level on the ramp is output in process step 6201, a short wait for settling is made (process step 6202). The corresponding position-indicator state variable $f(x)$ is then measured in process step 6203. The next discrete voltage level is then output in step 6201, unless a 'last step' decision is made in process step 6204; in which case the process ends with step 6205. Since a particular staircase signal is provided which is converted into the drive voltage V, and $f(x)$ is measured simultaneously, this in effect constitutes the outputting of $S(f^{-1}(f(x)))$, i.e. the functional dependence $S \circ f^{-1}$ of S upon f, where the circle symbol indicates function composition. The numerical value of the control parameter S used in the control loop 6100 is the transducer-coil current in voltage units – which is taken to be V. This procedure is approximately correct (in the case of a voltage controlled amplifier assumed here) to the extent that the non-Ohmic EMF terms in the coil circuit, including the effective coil inductance and BEMF voltage terms, are neglected. This is a justifiable approximation for sufficiently slow

ramping, i.e. long ramp-times and settling times. The ramp is made slow relative to audio signal timescales, because it is undesirable to put out audio information in the ramp. Therefore, the current into the coil is proportional to voltage by Ohm's law, to a good approximation.

However, care must be taken that the ramp not be too slow, for otherwise significant heating of the coil could take place, and the coil current through the coil would then drop due to increased coil resistance. Care must also be taken to minimize the thermal and viscoelastic hysteresis effects reflected in the staircase-ramping measurements. Additionally, what unavoidable hysteretic effects do remain~~,~~ should be compensated for via some averaging procedure. In preparing the curve of S as a function of x for an Audax 3" transducer, waveform 6206 shown in FIG. 22 included thirty-two steps of equal duration per each sweep from highest to lowest or lowest to highest voltage value. During the first and last of the steps the output voltage was zero. In each of the other steps, the voltage increment or decrement was $1/16th$ of the zero-to-peak amplitude of the waveform 6206, which was 0.25 volt. This value was before amplification. The amplitude of the ramp-sweep voltage signal fed to the voice coil of transducer 6103 was about 20 times higher. This amplitude is determined, for each speaker transducer, by the need to cover the full excursion of the coil/diaphragm motion that is encountered in normal operation.

Page 69, line 29, please amend the following paragraphs:

As for the B term in the control equation ~~(38)~~(40), measurement of the functional dependence of $B(x)$ upon $f(x)$, denoted as $B \circ f^{-1}(f(x))$, can be made by outputting a low amplitude tone, at a frequency sufficiently removed from the mechanical resonance frequency of the transducer to simplify the transducer's linear-response transfer function. The sound pressure output, or SPL, is measured at some fixed distance in front of the speaker, for example by means of a microphone, or alternately via other transducers within the speaker enclosure, or transducers in other speaker enclosures within a suitable proximity to the transducer being characterized. The off-resonance choice of tone frequency provides a relatively simple relation between the measured SPL and the motor factor Bl, which in turn is inversely related to B. The deduced values of B can then be tabulated against corresponding measurements of $f(x)$, for a stairway-ramped voltage signal 6206, in a manner similar to that used in the S measurements described above. At each DC voltage level, the low-~~low-~~amplitude tone is applied after that DC level has been held a sufficient time to allow electromechanical relaxation of the transducer to a steady state current and mechanical equilibrium. The frequency of the tone is fixed for each~~stairway-~~ stairway-ramped voltage sweep, but can be varied from sweep to sweep. However, the foregoing

approach is complicated by two factors. Firstly, the speaker's acoustic transfer function (diaphragm motion to SPL) is not a priori known for realistic speaker enclosures; and secondly, the suspension stiffness still affects the conversion of SPL values to B values, through the x_n-dependent elastic resonance frequency, for tone frequencies low enough so that coil-inductance effects do not spoil the simple Ohmic conversion of voltage to coil current. This latter fact means that the S and B measurements are effectively entangled, as the extraction of B values requires knowledge of S values; and the converse also holds, as explained below.

Because of these complications a hybrid approach is utilized, as follows. First, a Klippel GMBH laser-based metrology system is used to find an eighth-order polynomial fit to the function $Bl(x)$, and the ratio function

$$B(x) = \frac{Bl(0)}{Bl(x)} \tag{44}$$

where $x = 0$ is the equilibrium position, is computed and replaced with a suitable lower-order polynomial fit. Note that this initial stage need only be performed *once* per given speaker, since drifts in the motor-factor function $Bl(x)$ are almost entirely multiplicative, stemming from temperature dependence of the airgap magnetic field, and thus hardly affect the ratio $B(x)$. Next, a stairway-ramped voltage sweep of the type described above is performed, in which the position-indicator nonlinear state variable $f(x)$ and the actual position x are simultaneously measured. The latter is measured via a~~position sensor used with a Klippel GMBH metrology system. This~~ Klippel-type laser, which returns a voltage known to vary linearly with actual position to a high accuracy. And finally, the Klippel-derived polynomial fit to $B(x)$ is combined with the interpolated function $f(x)$ to yield an approximate polynomial interpolation for the composite functional relation $B \circ f^{-1}(f(x))$:

$$B \circ f^{-1}(f(x)) = b_0 + b_1 f(x) + b_2 f(x)^2 + b_3 f(x)^3 + \dots \tag{45}$$

Once interpolative approximations (polynomial or other) to both the functional relations $S \circ f^{-1}$ and $B \circ f^{-1}$ (i.e. both $S(x)$ and $B(x)$ as functions of $f(x)$) are determined, these interpolations are stored and integrated into the controller DSP and used, in transducer operation, to dynamically compute and output a corrected coil voltage V_{coil} from the original audio input signal w, via the control equation ~~(38)~~(40), as explained in Detail 10 below.

FIG. 23 is a general block diagram of a system 6300 depicting an audio transducer 6304 with the digital controller 6301. Digital controller 6301 received two inputs: the audio voltage signal w 6302 (also referred to as V_{audio}; see Detail 2), and the most recent measurement of the

position-indicator nonlinear state variable~~,~~ $f(x)$ 6303. This nonlinear state variable is measured in the transducer 6304. Digital controller 6301 combines the audio input with the measured value of $f(x)$ to compute the corrected V_{coil} in accordance to the control law. The control law may be that given by equation ~~(38)~~(40) in the event that only the transduction-process corrections are selected, or by other equations in Detail 2 in case the user decides to activate other combinations of control laws. The voltage V_{coil} is output in analog form 6305 by digital controller 6301, and provided to the amplifier 6306. The output voltage from amplifier 6306 is provided to transducer 6304.

As discussed in Detail 2, the use of the entire spring force in the correction, thus in effect electronically subtracting away the entire elastic restoring force, would lead to dynamical instability. It is therefore necessary to add back a linear spring restoring force calculated as an adjustable fraction of the measured spring factor at equilibrium, $S(0)$. This is done by subtracting a term linear in the estimated position $f^{-1}(f(x))$ from the ratio of the $S \circ f^{-1}(f(x))$ polynomial to the $B \circ f^{-1}(f(x))$ polynomial, since this ratio is the constant times an interpolating function for the suspension stiffness $x\,K(x)$. The net result of this subtraction is that the numerical values of S, and the functional relation $S \circ f^{-1}$, are replaced by new quantities, denoted here as S' and $S' \circ f^{-1}$ respectively, in the control equation ~~(37)~~(39). If the transconductance corrections are turned off, equations (36) and ~~(37)~~(39) reduce to the transduction-corrections equation ~~(38)~~(40), which is just equation (34), but with S replaced with the following subtracted value:

$$S' = S - k\,f^{-1}(f(x))\,B \tag{46}$$

~~Where~~ where ~~a~~ $k = q\,K(0)/R_{ms}$ is a constant multiplier~~,~~ related to the adjustable parameter q of equations ~~(37) and (38).~~(39) and (40). The multiplier q can be optimized by user preference. In Equation (46), the three quantities S, B and S' are all expressed as interpolated polynomials in the measured position- indicator nonlinear state variable $f(x)$, as described above.

Page 73, line 9, please amend the following paragraphs:

According to the present invention, there are ~~provided~~ parameterized linearization-filter functions characterizing the given transducer, which are measured and estimated using in-operation measurements of at least one nonlinear position-indicator state variable, augmented by preliminary (characterization) calibration runs in which this nonlinear state variable is measured

simultaneously with a more linear position-indicating variable (such as the Klippel-GMBH laser metrology system). The nonlinear position-indicator variable measured in operation can be a voltage output from an optical device, as is the case in one embodiment of the present invention and as is described in Details 8 and 13 below; or it could be an output from the internal electronic parameter measurements, as described in Details 6,7,11 and 12. These measurements could be augmented by an external measurement of sound pressure level during characterization runs, as described above.

Accordingly, ~~an invention where~~ the S and B parameters, which are needed by the controller to implement the transduction-process portion of the linearizing control law, can be matched to the program material by adjusting the parameter q governing the electronic spring force compensation, as described in equations ~~(37), (38)~~(39), (40) and (46).

DETAILED DESCRIPTION 6: Z_e MEASUREMENT THEORY.

An important aspect of the present invention is described in the context of a digital control system which linearizes audio reproduction using a position-indicator state variable, f(x), which is monotonic in position. The inductance of a transducer voice coil provides such a position-indicator state variable. ~~This method applies to many other classes of non-linear actuators and motors.~~

Page 74, line 11, please amend the following paragraphs:

It can be seen from curve 403 in FIG. 4 that the values of L_e (in this case at frequencies below $1\,kHz$) are monotonic with x; that is to say, no two distinct x values within the range $-2\,mm$ to $2\,mm$ correspond to the same value of L_e. We can thus map Bl (curve 401) and K (curve 402) onto L_e, and a measurement of L_e will uniquely predict both $Bl(L_e)$ and $K(L_e)$. These functional relationships are depicted in FIG. 24, in which curve 5101 is a plot of K in Newtons/mm and curve 5102 is a plot of Bl in Newtons/amp, both of which are plotted against L_e for the same data from FIG. 4. This new mapping provides a basis of a correction scheme. Because the inductance of the voice coil is a function of its position, by measuring the inductance the position of the voice coil is determined. Thus L_e provides an inductive position detector.

Page 77, line 9, please amend the following paragraphs:

A method for measuring the coil inductance is illustrated by the block diagram in FIG. 31. A supersonic probe tone ("carrier signal") is applied via input line 7401 to the voice coil of transducer 7402. In this approach, a reference $R\,L$ circuit 7403 is placed in series with the voice coil. The supersonic signal is then injected into the voice coil of the transducer 7402 in addition to the audio signal, and the voltage across the voice coil of the transducer 7402 and the reference $R\,L$ circuit 7403 is measured. Reference $R\,L$ circuit 7403 may be implemented using a resistor and a coil in series. Alternatively, a coil or a resistor may be used to implement circuit 7403. The measured voltage signals are sent via ~~summer~~ summing junction 7404 and ~~summer~~ summing junction 7405 through filter 7406 and filter 7407, respectively, and the ratio of the output of the filters is then determined in either the analog or digital domain. The filter 7406 and filter 7407 are band pass filters implemented about the frequency of the carrier signal. Envelope detection via envelope detector 7408 and envelope detector 7409 is used to extract the signal due to changes in L_e. The ratio of the voltages coming out of the envelope detector 7408 and detector 7409 can be described in the Laplace domain as:

$$V_{ratio} = \frac{L_e\, s + R_e'}{L_{ref}\, s + R_{ref}} \qquad (47)$$

~~Where~~where R_e' is the resistive component of coil impedance at the probe tone frequency, including both the Ohmic coil resistance R_e and the lossy effective coil impedance component due to eddy currents. R_{ref} and L_{ref} are the respective series resistance and inductance of the reference $R\,L$ circuit 7403; and s is the Laplace variable. Because the ratio of the two voltages is taken, the signals that are close in frequency to that of the carrier, and thus cannot be rejected by the band-pass filter 7406 and filter 7407, will not introduce significant error in L_e determination. As long as L_{ref} and R_{ref} are chosen so that $\frac{L_e}{L_{ref}}$ and $\frac{R_e}{R_{ref}}$ are the same for frequencies near the probe tone, V_{ratio} remains a constant equal to $\frac{R_e}{R_{ref}} = \frac{L_e}{L_{ref}}$, regardless of the presence of ~~-~~other signals in the system that are close to the frequency of the carrier signal. Since L_e varies with coil position x, V_{ratio} will change accordingly. FIG. 27 shows the Bode plot of the transfer function V_{ratio} given in equation (47), while FIG. 28 shows the corresponding phase Bode plots.

The ordinate in FIG. 27 is the magnitude of V_{ratio}, in dB units, while the ordinate in FIG. 28 is the phase of V_{ratio}, in degrees; in both plots, the abscissa represents angular frequency in

units of radians per second. In both FIG. 27 and FIG. 28, the family of Bode plots is for progressively larger values of L_e, with the highest L_e value resulting in the curve 7201 and curve 7204, while the lowest value results in the curves 7202 and 7205. It is seen that as L_e increases, so does the magnitude of V_{ratio}. The sensitivity of V_{ratio} to changes in L_e is clearly a function of the probe tone frequency. The higher this frequency, the more sensitive V_{ratio} will be to L_e variations.

To reduce the effect of the common mode in-band noise, which is present in the voltage across the voice coil (i.e. $(L_e s + R_e') \cdot i$) and ~~the~~ in the voltage across the reference $R\,L$ circuit (i.e. $(L_{ref}\, s + R_{ref}) \cdot i$), upon the voltage ratio, the phase shift of $\frac{L_e\, s + R_e'}{L_{ref}\, s + R_{ref}}$ must be small. Thus, the choice of probe tone frequency may have an impact on the effectiveness of noise cancellation within the above-described approach. Furthermore, to ~~insure~~ensure the noise cancellation advantage of this algorithm, the band pass filters, mentioned above, must be matched as closely as possible.

Page 79, line 22, please amend the following paragraph:

Because FIG. 27, FIG. 28, FIG. 29 and FIG. 30 illustrate that a thermal change in R_e is likely to be misinterpreted as a change in L_e, a modification in the algorithm is needed to separate this thermal effect from actual changes in L_e that are caused by changes in the voice coil position. From FIG. 29 and FIG. 30, it is clear that the effect of variations in R_e upon the ratio V_{ratio} is minimized at the higher probe tone frequencies. This characteristic of V_{ratio} can be utilized to accurately determine L_e in the presence of thermal changes to R_e. For instance, for the Labtec Spin 70 speaker transducer for which the curves in FIG. 27, FIG. 28, FIG. 29 and FIG. 30 were generated, the use of a carrier signal at $150\,kHz$ will significantly reduce the thermal effects upon L_e measurement.

DETAILED DESCRIPTION 7: C THEORY --- PARASITIC CAPACITANCE AND CANT DYNAMICS.

Page 80, line 8, please amend the following paragraphs:

An important aspect of the present invention is described in the context of a digital control system ~~which~~ that linearizes audio reproduction using a position-indicator state variable,

$f(x)$, which is monotonic in position. The parasitic capacitance $C_{parasitic}$ between the voice coil and the body of a transducer can be used to give such a position-indicator state variable. This method applies to many other classes of non-linear actuators and motors.

The parasitic capacitance $C_{parasitic}$ between the voice coil of a transducer and the body of the transducer is largely determined by the relative positions of the voice coil and the magnetic pole pieces and central core. The variation of this capacitance with position is relatively straightforward and robust (reproducible). As illustrated, for example in FIG. 3, typically the voice coil 303 fits about a central core -310 ~~which~~ that is part of the iron assembly 305. The variation in the parasitic capacitance depends largely on the overlap of the voice coil 303 with the central core 310, and, to some extent, with the outer pole piece 311 as well.

Page 81, line 4, please amend the following paragraph:

Importantly for the purpose of the present C method, $C_{parasitic}$ is an easily measurable internal circuit parameter of the transducer which is, at the same time, a state variable which depends monotonically upon axial coil position x. As the coil moves deeper into the magnetic airgap, the ~~capacitative~~capacitive contact areas between the metallic surfaces of coil and poles on the one hand, and between former and poles on the other hand, increases; and thus so does the value of the parasitic capacitance.

Page 83, line 2, please amend the following paragraphs:

It is possible to understand the results from FIG. 32 and FIG. 33 using simple semi-~~quantitive~~quantitative models. Although some fairly involved modeling is required to obtain an accurate prediction of $C_{parasitic}(x)$ for a given transducer, it is quite easy to estimate its order of magnitude. Thus, referring to FIG. 35, assume a coil of height h and radius r. In the modeling described below, it is assumed that the coil former is to be non-conducting; thus only the coil-slug contribution to the capacitance is considered. Furthermore, we ignore the capacitance between coil and outer pole (as the capacitive overlap area for that pair of conductors is assumed smaller than between coil and core). For simplicity the wire indentations and insulation are likewise ignored. The maximal value of $C_{parasitic}(x)$ occurs when x is smallest, that is to say, when the coil is farthest into the magnetic airgap, $x = x_{\min}$. Assuming that at this coil position the ~~capacitative~~capacitive contact area between coil and slug equals the total area of the coil's cylinder, the following estimate -results:

$$C_{parasitic}(x_{\min}) \approx \varepsilon_0 \, 2\pi r \, h / g_{interior} \qquad (48)$$

~~Where~~where ε_0 is the permittivity of air and $g_{interior}$ is an estimate of the average distance between the steel of the central pole, and the copper surface of a typical wire belonging to the coil's innermost winding layer. For instance, in the case of the ~~LabTec~~Labtec speaker transducer discussed above, the geometrical parameters are estimated to be $r = 7.5mm$, $h = 5mm$, and $g_{interior} \approx 0.2mm$. Substitution of these three values into equation (48) yields:

$$C_{parasitic}(x_{\min}) \approx 10pF \qquad (49)$$

The value measured electronically was found to be about $18pF$ for this transducer. The discrepancy is reasonable given the parameter estimates.

For transducers of smaller speakers, such as those utilized in cell phone receivers, smaller capacitance values, for example several picoFarads, were measured. This decreased magnitude can readily be understood from the way in which the right-hand side of equation (48) scales down with the linear dimensions of the speaker's transducer.

Page 84, line 18, please amend the following paragraph:

When an aluminum former is used as a heat sink for the voice coil, which is often the case in transducers of woofer speakers due to the high power levels dissipated in their coils, unwanted circumferential eddy currents are induced in the former. These eddy currents result from two effects: one is the EMF induced in the former due to the its axial motion through the radial magnetic field in the airgap; and the other is the EMF induced by the time dependence -of the coil-current's contribution to the axial magnetic field through the former's interior. In order to suppress these eddy currents, it is standard practice to interrupt them by introducing a slot along the axial length of the former's surface. This practice does not, however, completely eliminate the former eddy currents, but instead has the effect of distributing them nonuniformly around the former's circumference. These nonuniform currents, in conjunction with the static radial magnetic field in the airgap, cause magnetic Lorentz forces on the coil/diaphragm assembly ~~which~~ that lack azimuthal symmetry. These non-uniform forces lead to a non-vanishing torque, and therefore to canting. This former-caused canting effect is discussed in J. Vanderkooy, J. Audio Eng. Soc., Vol. 37, March 1989, pp.119-128.

Page 85, line 14, please amend the following paragraphs:

For the Labtec Spin70 transducer, canting due to fractional turns, in addition to exacerbating audio distortions, makes correction using the C method less desirable in some ranges of cone movement, by causing the function $C_{parasitic}(x)$ to become non-monotonic when

in operation. As the voice coil moves towards the back of the speaker through the airgap to, or beyond, its mechanical equilibrium point, the fractional wire-turns approach the region of high-magnetic-field in the airgap sufficiently to cause significant torque and canting; the cant, in turn, causes some parts of the coil wire's conducting surface to recede further from one or the other of the magnetic pole structures, increasing the value of the effective capacitive gap $g_{interior}$ in equation (43) and thereby *decreasing* the values of $C_{parasitic}$.

Page 86, line 14, please amend the following paragraphs:

It is assumed that the fractional part of the number of coil-wire windings is 1/2, and the above notation for coil dimensions is retained. A further simplification is made, in that the radial magnetic field at the position of the half-winding is replaced with the same field component averaged over all the coil's windings. The canting torque on the coil/diaphragm due to the magnetic Lorentz force, is then approximately:

$$\tau_{magnetic} \approx \frac{r}{2N} Bl(x)\; i(t) \qquad (50)$$

—~~Where~~where τ denotes torque; $i(t)$ is the coil current, time independent in the DC case; $Bl(x)$ is the transducer motor factor, and N the total number of windings in the voice coil.

This *magnetic* torque is opposed by an elastic torque, caused by the elastic restoring forces acting to counter the canting. We denote by $\frac{1}{4}h^2\, \rho_{elastic}(x)\, K(x)$ the relevant torsional spring constant – i.e. the elastic torque, per radian of tilt, exerted by the speaker's spider and surround upon the coil, diaphragm and cone; here h is the coil's height (defined above equation (48)), $K(x)$ is the coil/diaphragm suspension stiffness recognized in prior art, while $\rho_{elastic}(x)$ is a dimensionless elastic ratio modulus characteristic of the coil/diaphragm assembly. The $\rho_{elastic}$ ratio modulus is expected to be significantly larger than unity, as speaker diaphragms are designed to resist canting while allowing axial motion.

With the above definitions, the elastic restoring torque is simply:

$$\tau_{elastic} \approx \frac{h^2}{4} \rho_{elastic}(x)\, K(x)\, \theta(t) \qquad (51)$$

—where $\theta(t)$ represents the canting (or tilt) angle, in radian units, as a function of time.

When the coil is driven with a DC or quasi-DC current, mechanical equilibrium is attained when the magnetic and elastic torques balance: this occurs at a tilt angle of

$$\theta(t) \approx \frac{2r}{N\,h^2} \frac{Bl(x)}{\rho_{elastic}\, K(x)}\, i(t) \qquad (52)$$

Ignoring the coil-wire insulation, this tilt results in an increase in the parasitic capacitance, roughly estimated at:

$$\frac{1}{C_{parasitic}(x,\theta)} \approx \frac{1}{C_{parasitic}(x)} + \frac{|\theta(t)|}{16\varepsilon_0 \pi r} \tag{53}$$

~~Where~~where $|\theta(t)|$ is the absolute value of the tilt angle, and $C_{parasitic}(x)$ is the capacitance for the case of no canting.

Since the driven-coil measurements for the ~~LabTec~~Labtec speaker transducer were quantified in terms of coil-circuit *voltage* rather than coil *current,* we set $i(t) = V_{coil}(t)/R_e$ in the above equations, where R_e is the coil's Ohmic resistance (this relationship requires corrections in the AC case, as detailed elsewhere in this document-). Thus, for the DC case, equations (52)-(53) now yield the predicted fractional increase in- parasitic capacitance due to canting:

$$-\frac{\delta C_{parasitic}}{C_{parasitic}} \approx C_{parasitic}(x)\, V_{coil} \,\frac{1}{8\pi h^2 \varepsilon_0 R_e N}\, \frac{Bl(x)}{\rho_{elastic}(x)\, K(x)} \tag{54}$$

Page 88, line 16, please amend the following paragraphs:

This tilt angle would only result in a maximal lateral displacement of order $0.02\ mm$ for parts of the coil – too small to cause the coil to be physically blocked by the pole structure, but enough to result in discernible audio distortions. However, the estimate for the fractional change in stray capacitance is quite dramatic, and in agreement with the measurements made for this speaker transducer.

DETAILED DESCRIPTION 8: IR DIODE MEASUREMENT THEORY

Page 90, line 7, please amend the following paragraph:

The present invention is described in the context of controlling an audio transducer system in part by a system consisting of hardware and software.

Page 90, line 12, please amend the following paragraphs:

A DSP based controller 10101 consists of a ~~DPS~~DSP processor and software system 10102 and an interface system 10103 consisting of analog input/output and user interface software. Audio input is provided to DSP based controller 10101 through a signal-matching network 10104 ~~which~~that filters the audio input and provides the correct level of input to the interface system 10103. The audio input is acted on by the control routines in the DSP based controller 10101 and is output to a second signal-matching network 10105. The signal from the

signal_-matching network 10105 is provided to a power amplifier 10106. The output of power amplifier 10106 drives a speaker transducer 10107. A position sensor 10108, or sensors, is used to provide a position indication signal, indicating the position of the coil/diaphragm assembly of the speaker transducer 10107 to sensor signal conditioner 10109. Such position sensors could be, for example, the Z_e detector of Detail 6, or IR detector described in Details 8 and 13, or C detector described in Details 7 and 12. Sensor signal conditioning system 10109 is used to amplify and filter the positional signal and match it to the level required for the interface system 10103.

FIG. 39 is a block diagram of a particular embodiment of an audio reproduction system 15100 that includes a DSP based controller 10101. A Personal Computer (PC) 15101, which could be an eMachines T1742, is used as a control and user input environment for the DSP based controller 10101. The DSP based controller 10101 is implemented using a M67 DSP board 15102 and a A4D4 I/O board 15103 both manufactured by Innovative Integration Inc. (Simi Valley, CA). The M67 DSP board 15102 is a mother-board for the A4D4 I/O board 15103. The M67 DSP board 15102 contains a 106MHz TMS320C6701 floating point DSP manufactured by Texas Instruments and has been modified to add an inverter (74LS14) between JP14 pin 34 to JP23 pin 29. The A4D4 I/O board 15103 consists of four 16 bit analog-to-digital converters (ADCs) and four 16 bit digital-to-analog converters (DACs) with interface circuitry to the M67 DSP board 15102. A Lynx L22 card 15104 manufactured by Lynx Studio Technology, Inc (Newport Beach, CA) installed on the PC 15101 provides an audio signal 15105 ~~which~~ that is input to the A4D4 I/O board 15103. The Lynx L22 card 15104 receives input via Cool Edit Pro software 15106 (version 2) installed on PC 15101. The Cool Edit Pro software 15106 generates a '.wav' type digital sound file from a music source, which could be a CD player 15107 also installed on the PC 15101. After processing by the DSP based controller 10101, the corrected analog audio signal 15108 is output from the A4D4 I/O board 15103, and provided as an input to a 20:1 attenuator 15109. Output from the attenuator 15109 is provided as input to a Marchand PM224 amplifier 15110 with internal jumpers set to give a DC coupled amplifier. The Marchand PM224 amplifier 15110 is manufactured by Marchand Electronics Inc (Webster NY). The Marchand PM224 amplifier 15110 is used to drive a 3" transducer 15111 manufactured by Audax (Westlake Village, CA). The embodiment of audio reproduction system shown in FIG. 39 uses the IR method of position sensing. An IR detector 15112, the operation of which is described in Details 8 and 13, is used both to measure the position of the coil/diaphragm assembly of the 3" transducer 15111 and to match the signal to the input stage of the A4D4 I/O board 15103. The output 15113 of the IR detector 15112 is an input to the A4D4 I/O board

15103.

DETAILED DESCRIPTION 10: SOFTWARE AND PROCESS FLOW

Page 92, line 6, please amend the following paragraphs:

FIG. 40 shows the process flow used to linearize the transconductance component of the signal conditioning process and the transduction process of a given audio transducer, based upon the control model given by equations (36)-(~~37~~39) in Detail 2 above. FIG. 40 applies also for the case that only a subset of these corrections are applied.

In the process illustrated by FIG. 40, the first step 111001 entails measuring large signal (LS) transducer parameters. This step yields coefficients of polynomial interpolations for the functions $Bl(x)$ and $L_e(x)$. The measurements are performed using a Klippel GMBH laser metrology system, with procedure as detailed in Klippel System Manual dated May 2, 2002.

In a second step 111002, a software control program is invoked~~, for example the software control program in file 071119.txt included in the computer program listing appendix provided by the compact disks included with this application~~. In a third step 111003, the invoked software control program is run in 'Calibrate' mode in order to calibrate the functional relation between coil/diaphragm position x and the position-indicator nonlinear state variable, $f(x)$, which in one embodiment of the present invention is the voltage output of the IR circuitry: $x_{ir} = f(x)$. During this calibration, the software control program collects corresponding values of x as measured to an approximation by the Klippel laser, and $f(x)$, in relation to the corresponding values of voltage outputs as described in Detail 5 so that the dependence of $f(x)$ with x and the dependence of S with $f(x)$ can be determined.

An example of the software control program used in step 111003 is provided by FIG. 41, FIG. 42, FIG. 43, and FIG. 44. The data obtained from steps 111001 and 111003 are used to find Best Fit coefficients for lowest order polynomials of S, x, Bl and L_e as functions of x_{ir}, as indicated by step 111004. Here 'Best Fit' is defined as that curve which is of the lowest order and which does not exceed specified rms and maximum errors, subject to substantial weighting in the mid section of the range of the $f(x)$ variable. More details and specifics on 'Best Fit' are provided later in this section. The user then inserts the polynomial coefficients obtained from step 111004 into the Software Control Program – step 111005. Next, the user invokes the Software Control Program for Normal operation – step 111006 - and operates the program in Normal mode 111007.

FIG. 41 shows the structure of one embodiment of the Software Control Program that is used both for obtaining data during calibration 111003, and for operating in normal mode 111007 in which linearized sound is produced. The initialization process 111101 places the system in a known state. The software control system can then be selected to operate in calibration mode 111103, which consists of an S and an x calibration process, or to operate in the normal mode 111104. Typically, the first time around the user needs to select the calibration mode 111103, as indicated in 111003. After completion of calibration mode 111103, the system can be selected for normal operating mode 111104, in which the software controls the sound reproduction process through an Interrupt Service Routine (ISR) 111106. Note that the ISR functionality 111106 is also used in calibration mode. On an exit event 111105 prompted by the user, the system stops the program 111107. FIG. 45, FIG. 46, FIG. 47 and FIG. 48 cover the normal operation in detail, while FIG. 42, FIG. 43 and FIG. 44 cover the calibration mode in detail; all these figures are described later in this section.

Page 97, line 5, please amend the following paragraphs:

The operations within step 111305 are detailed in FIG. 51. In step 111401, for the given set of coefficients, for example, $c_0, c_1, \ldots\ c_9$ for a polynomial Y - values of Y_{orig} are calculated as follows:

$$Y_{orig}(p) = c_0 + c_1\,p + c_2\,p^2 + \ldots + c_9\,p^9$$

~~For~~for each of several points p in the range given above. The above $Y_{orig}(p)$ values are then used in step 111403 to compute new coefficients, and in module 111404 to compute errors. Here Y_{orig} values: Y_{orig1}, Y_{orig2}, $\ldots$ Y_{orig33} are calculated for 33 points $p1$, $\ldots$, $p33$ distributed uniformly over the above range. It will be readily recognized that that the number of points used can be changed within the framework of this invention.

Page 97, line 30, please amend the following paragraphs:

Let $z_j = w_j\,(Y_{orig})_j$-, $j = 1,\ldots,N$, be the weighted desired output vector, and b_k, $k = 0,\ldots,M$ be the reduced order vector of coefficients ~~which~~ that needs to be determined. Then the weighted output vector for the points p_j for the coefficient column vector b is given by the new column vector Ab. The total weighted squared error between the two weighted vectors is given by:

$$E = (z - Ab)^T\,(z - Ab)$$

Page 99, line 18, please amend the following paragraphs:

~~The Matlab program developed by Tymphany Corporation to implement module 111403, is named 'reduce_order_of_XlsrSB_Lcoeffs'. The code is included in the computer program listing appendix on the compact disks included with this application in files: 071115A.txt; 0711115B.txt; 0711115C.txt; 0711115D.txt; 0711115E.txt; 0711115F.txt; and 0711115G.txt. File 071115A.txt lists the main program itself, which calls the filenaming utility function 'fNmsInOutXlsrSB_Lcoeffs' (listed in file 0711115B.txt) for the purpose of letting user specify input and output filenames. Next, the main program calls the function 'reduce_order_of_coeffs', (file 0711115C.txt) for each of the 4 curves whose order is to be reduced. This function in turn calls 'WtdLstSqPolyCoeffs' (file 0711115D.txt), a subfunction that calculates the coefficients according to the weighted least squares approach described above in order to provide the best fit coefficients. In turn, the subfunction 'WtdLstSqPolyCoeffs' of file 0711115D.txt calls 'cnstrct_A_mat_a_z_col' (listed in file 0711115E.txt) to construct the weighted *A* matrix and the weighted *z* column vector needed for calculating the coefficients. The function 'reduce_order_of_coeffs' of file 0711115C.txt also calls a plotting routine, 'plt_data_sup_y3' (file 0711115F.txt), which plots curves if requested by user. Finally, the program calls the function 'wr1setRdcdCoeffToOpenFile' (listed in file 0711115G.txt) four times, in order to write one set of coefficients for each of the four curves.~~

Page 101, line 8, please amend the following paragraph:

FIG. 47 shows a flow diagram of the ISR 11303. When the negative edge of the sampling clock occurs, the software control passes from the wait loop and command parser 11204 to step 11401. Step 11401 limits the value of the word to be sent to the DAC 11402, so that it does not exceed the input range of the DAC 11402. The DAC can be an onboard DAC as it is with the Innovative Integration A4D4, or a serial-port-based off-board DAC. The analog signal that is created is the corrected audio signal V_{coil}, and is fed to a power amplifier 10106. To create the corrected audio sample, the ISR module 11303 uses IR sensor data $f(x)$ from module 11403 and audio data from module 11405. A digital filter 11404 is used to minimize sensor noise in the measurement of $f(x)$. Module 11406 computes S, B, and L_e corrections from the filtered value of $f(x)$ 11404, as described below.

Page 102, line 3, please amend the following paragraphs:

The corrected audio signal V_{coil}, calculated by a combination of actions by modules 11406, 11407 and 11408, is derived from input audio signal and the value of filtered $f(x)$ using the following eight equations:

$$Bl = Bl_0 + Bl_1\, f(x) + Bl_2\,(f(x))^2 + Bl_3\,(f(x))^3$$ (~~52~~57)

$$S = S_0 + S_1\, f(x) + S_2\,(f(x))^2 + \cdots + S_5\,(f(x))^5 - k\, f^{-1}(f(x))\,/\,Bl$$ (~~53~~58)

$$x_c = (x_c)_0 + (x_c)_1\, f(x) + (x_c)_2\,(f(x))^2 + (x_c)_3\,(f(x))^3$$ (~~54~~59)

$$L_e = L_0 + L_1\, f(x) + L_2\,(f(x))^2 + L_3\,(f(x))^3$$ (~~55~~60)

$$\hat{x}(t) = \alpha\, \hat{x}(t-\tau) + \beta\,(f^{-1}(f(x_c(t))) - f^{-1}(f(x_c(t-\tau))))$$ (~~56~~61)

$$\cancel{BEMF = (K_{V1}\, Bl - K_{V2}\,/\,Bl)\,\hat{x}(t)}\; BEMF = -(K_{V2}\,/\,Bl)\,\hat{x}(t)$$ (~~57~~62)

$$\cancel{V_1(t) = S + V_{audio}(t)\,/\,Bl + BEMF}\; V_1(t) = S + V_{audio}(t)\; Bl_0\,/\,Bl + BEMF$$

(~~58~~63)

$$\cancel{V_{coil}(t) = V_1(t) + (K_{I1}\, L_e - K_{I2}\, L_0)\,(V_1(t) - V_1(t-\tau))}$$

$$V_{coil}(t) = V_1(t) + K_{I1}\; L_e\;(V_1(t) - V_1(t-\tau)) + K_{V1}\; Bl\; \hat{x}(t)$$ (~~59~~64)

—where: V_{coil} is the corrected voltage signal applied across the voice coil and including all four corrections (S, B, BEMF and L_e); V_1 is the corrected voltage *without* the inductive correction; V_{audio} is the audio input voltage signal, suitably normalized; t and τ denote the current time-step and the sampling time, respectively; and the constant k in the subtraction term in the polynomial expansion for S (last term onright-hand side of equation ~~(53)~~(58)) is the electronic linear spring stiffness remaining after the linearizing filter (see Details 2 and 5 above). It is used in the calculation of S in order to maintain an appropriate level of restoring force in a transducer (see Detail 5 above); without this restoring term, the transducer would become unstable.

Equation ~~(54)~~(59) is a correction applied to linearize the IR position-indicator state variable $x_{ir} = f(x)$ if necessary. Equation ~~(55)~~(60) is the correction for nonlinear inductance L_e.

Equation ~~(56)~~(61) is a digital filter designed to estimate the velocity of the transducer needed for the BEMF correction. Equation ~~(57)~~(62) calculates the required BEMF correction. The BEMF correction comprises two components: the removal of the nonlinear BEMF and the replacement with a linear BEMF. The equations incorporate a multiplier for each term to allow for fine adjustment of the correction. Equation ~~(58)~~(63) and ~~(59)~~(64) implements the above components of the audio correction.

It will be appreciated that there are many different ways of discretizing the numerical differentiation operation of the control diagrams FIG. 11 and FIG. 12, and that the

implementation of these numerical differentiations used in one embodiment of the invention, and shown in equations ~~(56)~~(61) and ~~(59)~~(64), represent but one possible choice.

Digital filters may be added to equation ~~(59)~~(64) for smoothing, equalizing and noise reduction. The polynomial coefficients as well as the powers of filtered $f(x)$ are stored in arrays, so that the needed sum of products can be easily computed. Moreover, the array for powers of filtered $f(x)$ may be constructed recursively, again reducing the computational cost.

Page 103, line 25, please amend the following paragraphs:

FIG. 42, FIG. 43 and FIG. 44 show flow diagrams of S and x versus $f(x)$ calibration 111103.

For calibration, the mainline loop is finite (while that in- normal mode is infinite) and results in a tabulated output, from which a polynomial curve is fitted and polynomial coefficients extracted for use in the Normal Mode 111104.

Page 104, line 7, please amend the following paragraphs:

FIG. 43 shows the details of HW and ISR operations for S calibration 11504. It depicts Sampling Clock 11601 and ADC Convert 11602, which are similar to corresponding modules in FIG. ~~113~~46; the same description applies, and is thus not repeated. Modules 11604 and 11605 limit and convert the digital values to analog waveform. Module 11606 tests whether the data is to be collected. During calibration mode, the mainline S calibration loop 11505, detailed below in FIG. 44, sets and clears the flag 'Collect_data'. If this flag is set, the data collection ~~proceeds as~~is done by the module 11607, and a sample count is tallied. Also, module 11608 reads the S value from the array, to be used in the variable 'dacvalue'. If the flag is not set, these two modules are bypassed. Module 11609 executes the return from ISR.

FIG. 44 shows the details of mainline S calibration loop 11505. Module 11701 checks whether any value of S is left with which to operate the loop. If there is one, it executes the path comprising modules 11702 through 11707 to send out the S value via the ISR 11603, and to collect the corresponding value of $f(x)$ and x as follows. Module 11702 executes a wait of 100 milliseconds to allow the transients in the transducer to attenuate. Module 11703 sets the 'Collect_data' flag which signals the ISR 11603 to collect data. Module 11704 allows 1 millisecond to collect samples, which at 48 kSPS collects 48 samples. These samples suffice to give a good reading of $f(x)$, the IR data, and x, the laser data. Module 11706 performs averaging, and the module 11707 stores S, $f(x)$ and x for offline curve fitting. As long as there is an S value to be covered, the process continues.

Page 105, line 17, please amend the following paragraphs:

FIG. 48 illustrates the details of the Wait Loop and Command Parser 11204, shown in FIG. 45, which is abbreviated below as WLCP. The system enters into the 11801 step of WLCP from Enable ISR Setup and Enable Sampling Clock 11202; in step 11801 it is determined whether Normal Mode operation should stop. If 'Yes', system enters into step 11803, in which Interrupt is disabled and the HW is put into a known state; then system is passed out of WLCP and into User Mode Select 111102. But if the answer to the 'Stop?' query (step 11801) is 'No', the DSP passes to step 'Command?' 11802, in which the WLCP checks to see whether User has entered a keyboard command since the last check (checks are spaced several microseconds apart during the Wait Loop). If no new keyboard command has been entered during the most recent such time interval, this is interpreted as a 'No' response to the 'Command?' query-, and the system is looped back to this 'Command?' query 11802. But if and when WLCP finds that a new keyboard command *has* been entered during the most recent time interval, each of the following optional keyboard responses are interpreted by WLCP as a 'Yes' and acted upon. User keyboard response 'c' causes the DSP to begin implementing corrections: 'Corrected Audio Mode' 11804; after this mode is entered, the system is passed back to the 'Stop?' query 11801. User keyboard response 'b' causes the DSP to enter the mode 'Adjust Linear BEMF' 11805, from which it is again returned to 'Stop?' query 11801. The following are the remainder of the allowed keyboard responses, and their effects. Response '+' puts the DSP into mode 'Increase Volume' 11806, from which it returns to 'Stop?' query 11801; similarly, response '-' puts DSP into mode 'Decrease Volume' 11809, and thence to 'Stop?' query 11801. Response 'u' puts DSP into 'Uncorrected Audio Mode' 11807, and thence to 'Stop?' query 11801. Response 'i' puts DSP into mode 'Adjust dL/dx Correction' 11808, and thence to 'Stop?' query 11801. Response 'o' puts DSP into mode 'Adjust Offset' 11810, and thence to 'Stop?' query 11801. Response 'j' puts DSP into mode 'Adjust dL/dx Offset' 11811, and thence to 'Stop?' query 11801. Response 'm' puts DSP into mode 'Mute On' 11812, and thence to 'Stop?' query 11801. Response 'k' puts DSP into mode 'Adjust Linear Spring' 11813, and thence to 'Stop?' query 11801. Response 'f' puts DSP into mode 'Turn IR Filter On' 11814, and thence to 'Stop?' query 11801. Response 'n' puts DSP into mode 'Mute Off' 11815, and thence to 'Stop?' query 11801. Response 'v' puts the DSP into mode 'Adjust Nonlinear BEMF' 11816, and thence to 'Stop?' query 11801. Response 'd' puts DSP into mode 'Turn IR Filter Off' 11817, and thence to 'Stop?' query 11801. And finally, a User response 's' puts the DSP into 'Stop' mode 11818, from whence the system is returned to 'Stop?' query 11801. It should be noted that all processes within the Wait Loop and

Command Parser, are interruptible by ISR 11303. ~~The C programming language code implemented in the DSP is provided in file 071119.txt on the compact disks of the computer program listing which is a part of this application.~~

DETAILED DESCRIPTION 11: Z_e METHODS AND CIRCUITS

Page 107, line 7, please amend the following paragraph:

In particular, the position-indicator state variable~~,~~ $f(x)$~~,~~ utilized in this embodiment of the invention is an output voltage derived from the functional dependence of the effective complex coil impedance $Z_e(\omega, x)$ upon coil/diaphragm position x, at some fixed supersonic probe frequency ω. The physical effects ~~which~~ that give rise to this functional dependence, along with a mathematical model developed to simulate them, in accordance with the present invention, are described in Details 1 and 6. This embodiment is called the Z_e method.~~;~~ ~~in~~ In this section we elaborate on the methods and circuits used to implement the Z_e method.

Page 108, line 4, please amend the following paragraph:

In the context of an audio transducer, the input signal to the voice coil will include audio information (*program material*) together with the probe tone. It is therefore necessary to separate the probe tone and program material in frequency, so that the probe tone measurement is not interfered with by the audio drive signal. The Nyquist ~~criteria~~criterion suggests that the probe tone 12101 should have a frequency of at least twice the audio frequency bandwidth, to avoid aliasing with the program material. A probe tone having a frequency of 43kHz has been found to be particularly desirable. However, many other frequency values could be used.

Page 109, line 18, please amend the following paragraph:

FIG. 55 shows a circuit diagram of the summing circuit 12202. The audio drive signal 12201 is provided as an input to filter 12401 which isolates the probe tone 12101 from the low impedance of the audio amplifier output. The filter 12401 is composed of resistive, ~~capacitative~~capacitive, and inductive elements, as indicated in FIG. 55. The probe tone 12101 is provided to a capacitor 124C4, which in turn is connected to the summing point 12402. Capacitor 124C4 decouples the audio drive signal at the summing point 12402 from the low impedance output of the sine wave generator 12304. The signal at the summing point 12402 is provided at output terminal 12403 which is connected to an input of the potential divider circuit 12203.

Page 110, line 21, please amend the following paragraph:

FIG. 58 shows the circuit of the low pass filter 12206. The first part of the low pass filter, incorporating the operational amplifier 127OA1, is a standard 2nd order Butterworth low pass filter. The second part of the filter is an inverting amplifying stage ~~which~~ that includes an operational amplifier 127OA2 and a variable resistance 127VR1 that produces a DC offset in the output signal. This offset is set to reduce the DC offset in the magnitude of the probe tone that has been detected. The gain of the inverting amplifying stage is set to enhance the signal significance when it is converted to digital form. One skilled in the art will recognize that many different circuit arrangements could be used for the filter, gain and offset circuit, and that the rather straightforward circuit shown in FIG. 58 can be modified without changing the essence of the design. Operational amplifiers 1270A1 and 1270A2 may be National Semiconductor part number LM741.

Page 111, line 18, please amend the following paragraphs:

The filter based method used in the- $Z_e(x)$ detection circuit 12200 and shown in FIG. 54, is sensitive to changes in output impedance of the audio amplifier 12303. For example, with a low impedance load, some types of amplifiers exhibit large crossover distortion effects, which in effect are a change in output impedance. This change in output impedance can cause noise in the $Z_e(x)$ measurement. Furthermore, in transducers driven with large currents there can be considerable heating effects in the coil. This produces a change in the Ohmic resistance R_e ~~which~~that is misinterpreted by the $Z_e(x)$ detection circuit 12200 as a change in position (this is discussed in Detail 6 above). Someone skilled in the art would recognize that a more complex circuit is required to separate out these two effects for the full range of transducers, but that this would not materially change the invention detailed here.

It will be apparent to those skilled in the art that the particular position-indicator state variable $f(x)$ described in this section and in Detail 6, which is derived from the functional dependence of the effective complex coil impedance $Z_e(\omega, x)$ upon coil/diaphragm position x at some fixed supersonic probe frequency ω, can be used within various embodiments of a feedback linearization control system according to the present invention, in which the positional information $f(x)$ is used in various different ways, including but not limited to one or more of the control laws presented in Details 2 and 10 above.

Page 112, line 27, please amend the following paragraphs:

FIG. 34 shows a schematic cross section of a typical cell phone speaker or receiver 13100; actual three-dimensional speaker geometry is a figure of revolution about the central horizontal axis of symmetry (not shown). Speaker 13100 consists of a transducer and integral acoustic venting. A voice coil 13101 is mounted on the diaphragm 13102. Coil 13101 is positioned in the gap between a neodymium magnet 13103 and a magnetic base plate 13104. A plastic surround 13105 supports the diaphragm 13102 and a ~~face-plate~~faceplate 13106. The surround and ~~face-plate~~faceplate have acoustic vents 13107 which tune the frequency response of the speaker 13100. The depth, indicated in FIG. 34 by D1, is typically 2 mm. The main difference between this type of transducer assembly and the transducer shown in FIG. 3 is the single surrounding support of the relatively flat diaphragm 13102. This means that the system is resistant to the tilt ("canting") ~~which~~ that can complicate capacitance position-sensing methods in other transducers as described in Detail 7.

The preferred method of detecting the variation of capacitance with coil/diaphragm axial position, $C_{parasitic}(x)$, is to place the capacitance within an oscillator circuit. Changes in $C_{parasitic}(x)$ due to changes in coil position are then the cause of changes in the oscillator frequency. A frequency-to-voltage converter is then used to yield a varying signal which is a function of the parasitic capacitance. The varying signal can be identified with $C_{parasitic}(x)$ in suitable units. Thus, as defined, $C_{parasitic}(x)$ can be identified with the position-indicator state variable~~,~~ $f(x)$.~~,~~

FIG. 60 shows a schematic of the capacitance detector and speaker arrangement, together with the DSP used for correction. An analog audio signal, provided over input line 13201, is digitized by DSP based mixed-signal controller 13202. Mixed-signal controller 13202 is embodied by a AD21992 chip which includes an ADC (analog ~~voltage~~ to digital~~)~~ converter). The output of the DSP based controller 13202 is connected to a standard DAC (digital to analog ~~voltage)~~ converter) 13203. The output of the DAC 13203 is amplified by a DC–connected audio amplifier 13204. The output of amplifier 13204 has a drive connection 13205a to one terminal of the voice coil 13101 of the speaker 13100. The magnetic base plate 13104 of the speaker 13100 has a connection 13207 to one input of an oscillator circuit 13208 (detailed in FIG. 61). Another input to the oscillator circuit 13208 is connected to the drive connections 13205a and 13205b of the coil 13101 through blocking capacitors 13209a and 13209b, respectively. The output of oscillator circuit 13208 is connected to a frequency to voltage converter 13210, which converts the variable frequency received from the oscillator circuit 13208, and also amplifies and level-

shifts the varying voltage output. The output 13404 from the frequency to voltage converter 13210, which is a measure of $C_{parasitic}(x)$ (abbreviated as $C_p(x)$ in the Figure), and hence the position-indicator state variable~~,~~ $f(x)$, is input into the mixed signal DSP controller 13202. Inside DSP 13202, both the analog output voltage from 13210 and the analog input audio signal 13201 are converted into digital signals, and combined by the DSP 13202 to yield the digital output 13211 of the DSP 13202. The purpose of the DSP functionality within the controller 13202 is to furnish the DAC 13203 with a digital signal such that the output of DAC 13203, after amplification by amplifier 13204, will feed the speaker-transducer voice coil with a voltage signal including both the audio program and a pre-distortion calculated to cancel out a significant portion of the nonlinearities introduced by the transducer in the course of its normal uncorrected operation. ~~This effect of the position-sensor analog signal 13404 fed by frequency to voltage converter 13210 into the mixed signal controller 13202, is termed feedback linearization.~~

Page 115, line 17, please amend the following paragraphs:

FIG. 62 shows the detailed circuitry of the frequency to voltage converter 13210. Frequency to voltage converter 13210 consists of two parts: a frequency to pulse converter circuit 13401, and a low pass filter, amplifier and level shifter circuit 13402. The frequency to pulse converter 13401 consists of a mono-stable multi-vibrator circuit 13407 ~~which~~ that includes an industry standard multi-vibrator ~~which~~ that may be, for example, a 74LS123 as used in this embodiment. The mono-stable multi-vibrator circuit 13407 takes the square wave output signal 13304 received from the oscillator circuit 13208, which has a constant rms value, and converts it to a pulse train ~~which~~ that is provided on line 13403. The pulse train 13403 has an rms value varying with frequency, which is a function of the transducer coil/core capacitance C_p, which in turn varies with coil/diaphragm position x. The low pass filter, amplifier and level shifter circuit 13402 converts the pulse train on line 13403 to a varying analog voltage output provided on line 13404. This varying analog voltage on line 13404 represents the varying capacitance $C_p(x)$. The low pass filter, amplifier and level shifter circuit 13402 includes an operational amplifier 1340A1, which receives the output signal on line 13403 and, using a gain of 10 as determined by resistor values, low-pass-filters and offsets the signal 13403; and operational amplifier 1340A2, which has a gain of unity and implements a second-order Butterworth filter. These operational amplifiers may be embodied, for example, as National Semiconductor part number LM741, or equivalent. Resistor 134VR1 is adjusted such that the coil/diaphragm equilibrium position produces a zero output voltage. Operational amplifier 1340A2 receives, at its input terminal 13406, the output signal provided at output terminal 13405 of operational amplifier 1340A1, and

then converts that signal to a voltage which is provided on line 13404 to mixed signal DSP 13202.

Page 116, line 30, please amend the following paragraphs:

The particular position-indicator state variable $f(x)$ described in Detail 7 and in this section, which is derived from the internal parasitic capacitance $C_{parasitic}$ between the transducer voice coil and the transducer magnetic pole structure, can be used with various embodiments of a feedback linearization control system according to the present invention, in which the positional information $f(x)$ is used in various different ways, including but not limited to one or more of the control laws presented in Details 2 and 10 above.

Page 119, line 6, please amend the following paragraphs:

The IR-LED 14201 and PIN diode 14202 are epoxied side-by-side onto the transducer frame 14203, with both diodes pointing at a reflecting region 14204 on the transducer cone 14205. Reflecting region 14204 should subtend a sufficient angle such that, as the transducer cone moves, the PIN diode 14202 detector admittance cone is always pointed within the region. The diodes are preferably inclined towards each other and pointed towards the axis of the transducer at approximately a right angle to the direction of motion, or towards the curve of the cone. As was noted in Detail 8 above, the PIN diode output is not completely linear with cone position and therefore requires calibration by comparison with a metrology system. The position–indicator variable, $f(x)$, and the degree of its non-linearity, can be varied by changing the positions and orientations of the two diodes relative to each other and to the transducer cone. Thus, there is some variation from one implementation to another and some adjustment by trial and error may be necessary.

Page 120, line 3, please amend the following paragraphs:

The following sequence of adjustments is iterated five to seven times, until the output waveform 14401 is about 90% of peak A/D limit:

(a) Increase IR-LED current by adjusting variable resistor 144VR1, and thus output power, until the magnitude of output signal 14401 is at the limit on one excursion;

(b) Adjust the balance by changing variable resistor 144VR2 until there is no output signal at terminal 14401;

(c) Adjust the gain of amplifier 1440P2 using variable resistor 144VR3 for desired peak-to-peak voltage corresponding to full motion of the transducer cone;

(d) Turn off the coil current, readjust the balance using variable resistor 144VR2, and zero the signal 14401 when the transducer cone is at the equilibrium point.

DETAILED DESCRIPTION 14: IR RESULTS

Page 120, line 24, please amend the following paragraphs:

Audio distortions were measured, both with and without the correction, by applying an industry-standard two-tone SMPTE test, with audio input consisting (instead of the CD player) of a $60\,Hz$ tone in conjunction with a $3\ kHz$ tone. All four corrections described in Detail 2 were applied by the DSP-~~-~~based controller: transducer correction (spring correction S and motor factor correction B), the BEMF correction, and the position dependent inductive correction.

FIG. 65 shows a portion near $3\ kHz$ of the FFT power spectrum distribution of the SPL (sound pressure level) wave-pattern picked up by a microphone in the acoustic near-field; both corrected spectra which is indicated by reference character 1521, and uncorrected spectra which is indicated by reference character 1522 are shown, and it is clearly seen that the powers in the $60\,Hz$-spaced lattice of intermodulation frequency peaks, are significantly reduced when the correction is applied. FIG. 66 shows the low-frequency portion of the same power spectrum distribution, showing multiple harmonics of the $60\,Hz$ tone; again, spectra are depicted both with and without correction, and again, significant reduction in the magnitude of the harmonic distortion peaks can be seen.